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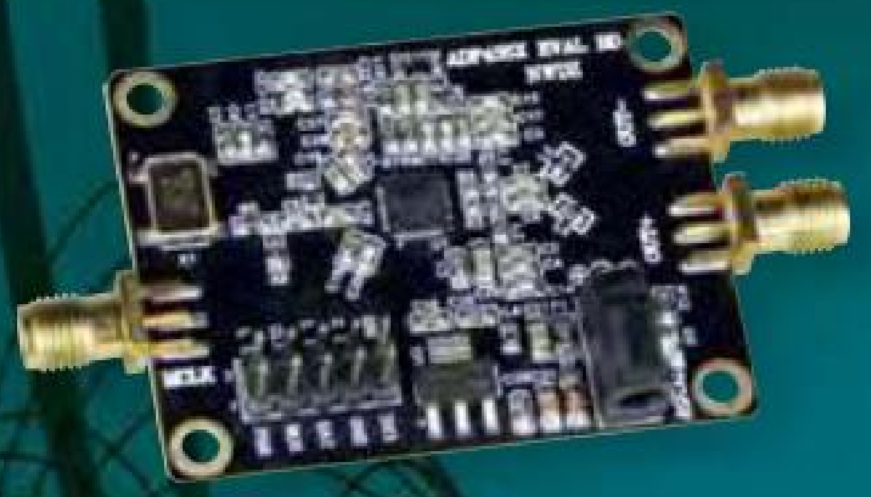

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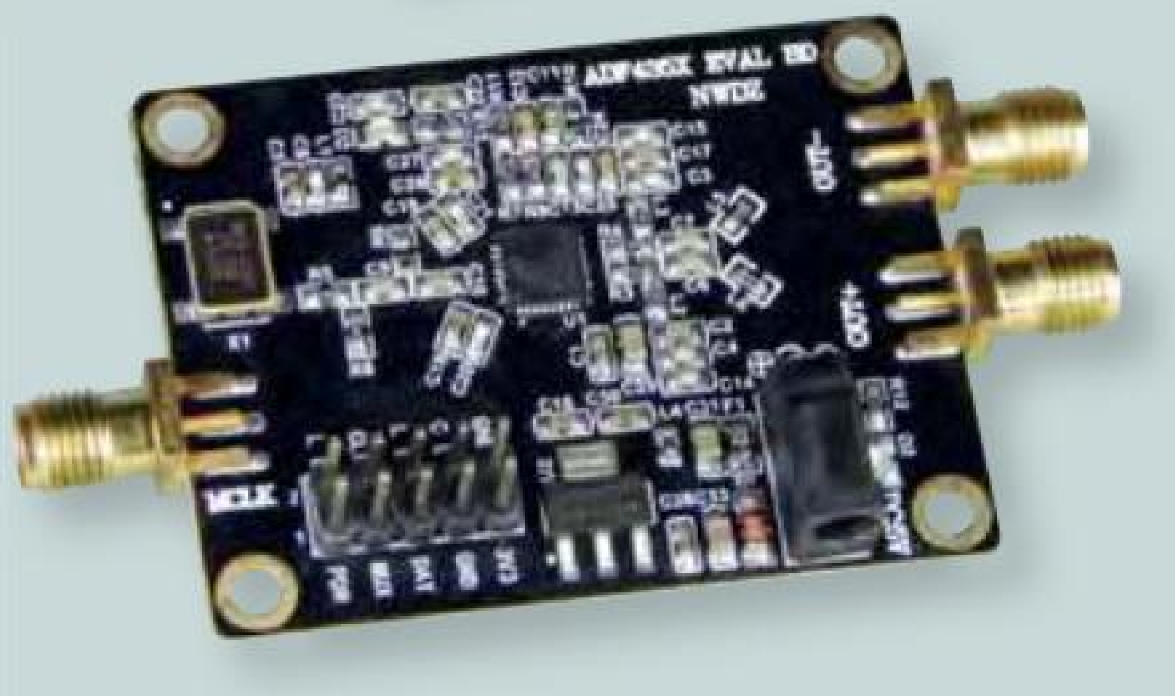

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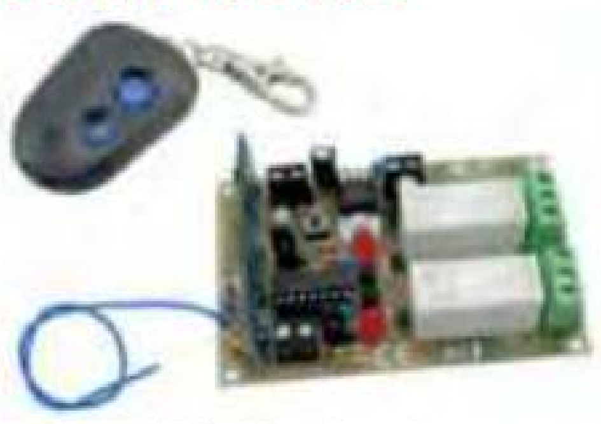

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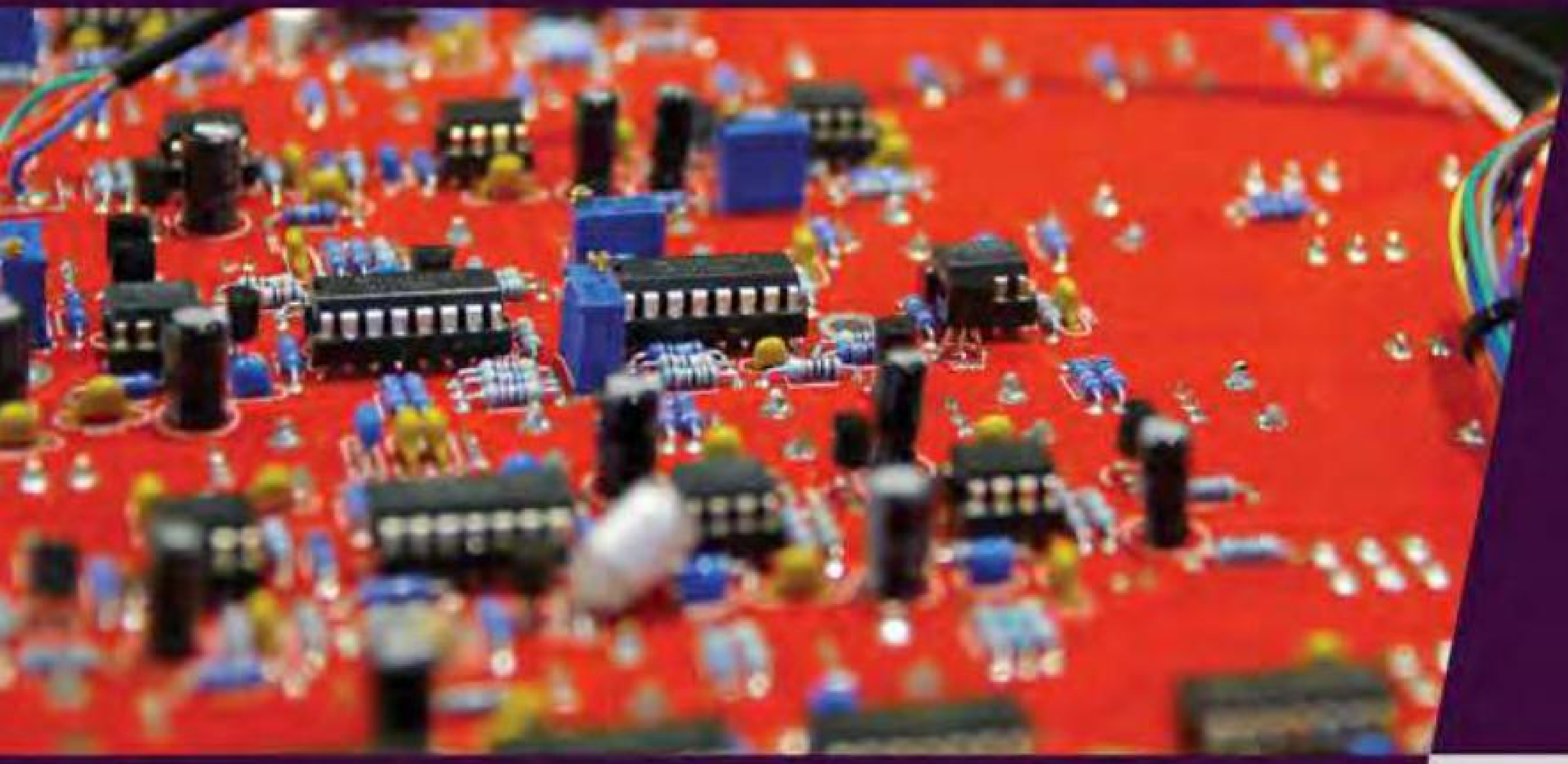

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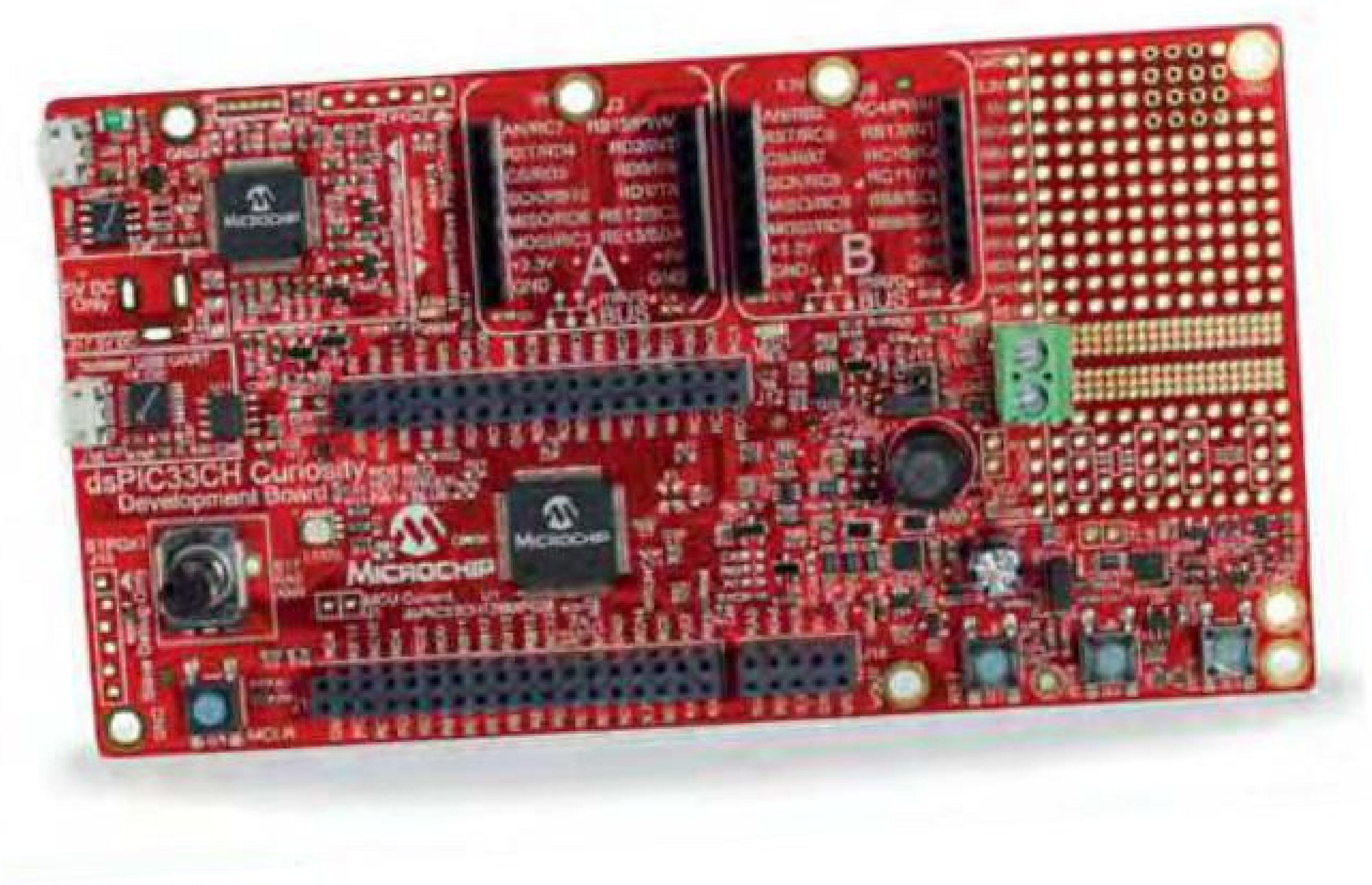

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Phew!

Welcome to the second issue of *Practical Electronics*.

Renaming and redesigning a popular magazine is always something to be undertaken carefully. On the one hand you don't want to alienate a loyal existing readership – but on the other, if improvements are needed then there comes a point when you just have to take a leap of faith, do your best, and ultimately… hope for the best.

Waiting for *PE's* first sales figures from our distributor was a little nerve wracking, but I'm pleased to say that it sold more copies than April 2018, which is particularly reassuring because that was last year's best seller.

Also, we appreciate the positive messages of support that you have sent in. It seems that for many of our long-term subscribers the revival of the name *Practical Electronics* has brought back some fond memories – thank you John E and Stephen A for taking the time to write.

While positive feedback is of course welcome, we also need constructive 'negative' feedback. Any observations, tips or requests help us give you a better magazine. Do write in.

So, from all of us here at *Practical Electronics*, very many thanks for maintaining your faith in the magazine.

Packed issue

What do we have for you this month? – a lot of goodies. There are four standalone projects to keep you busy – I particularly recommend the *USB Port Protector* if you are in the habit of powering projects from your computer. As Nicholas Vinen, the project's author puts it, 'Don't let one small "oops" fry your PC!'. Wise words, and not ones you want to learn after the event.

Phil Boyce's *Make it with Micromite* microcontroller series is gently accelerating to the point where you can start to read inputs and generate outputs. For those of you – especially beginners – who have not been following the series, I strongly recommend you reconsider. It is a fabulous route to very-low-cost, easy-to-implement and hugely satisfying intelligent electronic control.

For those of you who want truly tiny control systems then Mike Hibbett's latest *PIC n' Mix* is just what you've been waiting for – a new series on the PIC10F202 microcontroller.

Next up, we have Paul Cooper's *MIDI Ultimate Synthesiser*. Since I'm not a musician, I must confess that I was really only dimly aware of what a truly professional-level analogue synthesiser could achieve. That all changed when Paul sent me a YouTube link (see article) demonstrating the *MIDI Ultimate*. It's impressive, and I hope you'll watch the video and consider building your own.

There's more, much more… but I'm out of space, so I'll hand over to you to enjoy this second issue of *Practical Electronics*.

Matt Pulzer
Publisher

The future of TV

4K, WCG, 8K, OETF… Barry Fox disentangles the latest acronyms, gets a handle on the future of TV and talks to the BBC tech gurus who explain that it is not just about pixel count.

The International Moving Image Society's (IMIS) 88th Annual Awards Ceremony and lecture held recently in memory of Bernard Happé (once technical manager of Technicolor and author of seminal books on movie film technology) featured two BBC top tech gurus sharing the inside story of the joint work done by the BBC and Sky on shooting the 2018 British Royal Wedding in 4K UHD HDR, and the BBC's work with 4K UHD on sporting events including the 2018 FIFA World Cup and Wimbledon Tennis.

Phil Layton (head of broadcast and connected systems, BBC R&D) kicked off with his views on the need – or otherwise – for 4K resolution, with 8K coming soon.

'Screen sizes of 50/55 inches are needed to show the benefits of 4K UHD, at the average UK viewing distance of 2.6m' he said. 'This holds good for screen sizes up to 100 or 110 inches. So 8K is only needed if screens get larger than that or people start moving closer'.

Andrew Cotton (principal technologist, BBC R&D's Broadcast & Connected Systems Section) offered some basic background on HDR (high dynamic range) video, with WCG (wide colour gamut).

Signal sensitivity

The human eye's sensitivity to light is approximately logarithmic, with a dynamic range of around 46 stops; but the eye uses a moving range window of around 14 stops. Modern digital cameras capture 12-14 stops, and when TV signals are distributed with 8-bit code the dynamic range is around 5-6 stops.

HDR seeks to bridge the gap between human vision and digital capture and display. The camera sensor uses an OETF curve (opto-electrical transfer function) and the display uses an EOTF curve (electro-optical transfer function). The OETF and EOTF curves are not same, so an OOTF processor (opto-optical transfer function) is used to compensate.

The HLG (hybrid log-gamma) HDR system developed by the BBC and Japanese broadcaster NHK, adapts to whatever the screen can handle. Dolby's rival proprietary system DolbyVision relies on PQ (perceptual quantisation) and continually changing metadata to control the display. Manufacturers must pay Dolby a royalty to use DolbyVision. Hardly surprisingly, the BBC uses only HLG, which needs no such metadata and is royalty free.

Andrew Cotton then explained the practical challenges of using any HDR system: 'The challenge for HDR is live sports, where there can be 40 cameras, most of them different – including stump cams inside cricket stumps – and a mix of SDR (standard dynamic range) and WCG/HDR is inevitable. For the Royal Wedding we were using 76 cameras. I had to go out to a local shop and buy a memory card so that we could download and install some new software updates.

'For the foreseeable future mixing SDR and WCG/HDR camera signals will be inevitable, and conversion techniques developed by the movie industry do not work well for TV. Fortunately, most UHD HDR cameras also output an HD signal. And the BBC has developed 'scene light conversion' software to match HLG HDR and SDR signals. Details to follow….'

'There is still work to be done, though, especially on matching colour, even when cameras output both BT.709 colour for SDR, and BT.200 for HLG HDR. We thought

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we were doing all right at the Royal Wedding until Chris Johns (Sky's chief engineer) pulled out a Coke can and we found the reds from two makes of camera were not matching. Comparing camera outputs is like comparing phone tariffs!

'For the World Cup we found that HBS (host broadcast services), which provides the broadcast facilities for FIFA, use a proprietary HDR system from Sony, called S-Log3. So we had to convert that, too. And we had to take account of the fact that the grass at the Wimbledon tennis grounds actually has a blue tinge.'

Streaming

Said Phil Layton: 'Because the BBC does not have any chance of broadcasting UHD by terrestrial TV, we could only distribute the pictures by streaming with iPlayer, using HEVC/H.265, Main 10, Level 5.1 compression. However, TV works with frames per second; for example, iPlayer works with four-second chunks of audio and video data. We stream four different quality levels, ranging from HD up to UHD, with adaptive bit rates of between 1.4 Mbps, 2.5 Mbps and 22.7 Mbps, and the receiver continually checks the available speed and self-adjusts.

'We streamed a soccer event on 17 February 2018 and everything was fine; we streamed gymnastics on 8 April and it was a disaster, everything that could go wrong did go wrong. But we were learning all the time. The good thing about streaming is that the iPlayer system knows exactly what is happening and when. For the World Cup and Wimbledon we were serving 400,000 devices, with data peaking at 1.7 TBit/s. There were 1.6 million stream starts. We had warned that we would have to cap the number of UHD streams at 60,000 to protect the HD stream, but we only had to do that once during the World Cup. The number of UHD streams was usually running at around 40,000. The analysis graphs show steep peaks when the commentator reminded viewers that they could stream in UHD. On only one occasion the network broke and we had to shed all viewers.

Consumer choices and quality

'Possibly the most significant thing we learned was that not all TVs and not all ISPs are equal. Quality of service depends heavily on the device. We could see that viewers were mainly using eight types of receiver and five main ISPs. From 3000 completed viewer surveys we could see a wide range of opinions on picture quality, with phrases like 'dull pictures' and 'lacking vibrancy'. Consumer equipment has very variable performance. We can't reveal the names of the TVs and ISPs that came out best and worst, so what we have to do now is find a way of explaining to viewers that what they see depends very much on what make of TV they are watching and what ISP they are using. How to do that is an on-going discussion.'

'But we proved that streaming UHD by IP at scale is possible.

'The one thing that really stands out is that people hate the latency that means they hear neighbours who are watching live, shouting a minute or more before they see and hear from a stream. We think we have a solution, which we are now trialling.'

Barry Fox is a Fellow of the International Moving Image Society

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Marconi 2955A	Radio Communications Test Set	£725
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Marconi 6200A	Microwave Test Set 10MHz – 20GHz	£1,950
Marconi 6200B	Microwave Test Set	£2,300
Marconi 6960B	Power Meter with 6910 sensor	£295
Tektronix TDS3052B	Oscilloscope 500MHz 2.5GS/s	£1,250
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Tektronix TDS3012	Oscilloscope 2 Channel 100MHz 1.25GS/s	£450
Tektronix 2430A	Oscilloscope Dual Trace 150MHz 100MS/s	£350
Tektronix 2465B	Oscilloscope 4 Channel 400MHz	£600
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Farnell LF1	Sine/Sq Oscillator 10Hz – 1MHz	£45
Racal 1991	Counter/Timer 160MHz 9 Digit	£150
Racal 2101	Counter 20GHz LED	£295
Racal 9300	True RMS Millivoltmeter 5Hz – 20MHz etc	£45
Racal 9300B	As 9300	£75
Solartron 7150/PLUS	6½ Digit DMM True RMS IEEE	£65/£75
Solatron 1253	Gain Phase Analyser 1mHz – 20kHz	£600
Solartron SI 1255	HF Frequency Response Analyser	POA
Tasakago TM035-2	PSU 0-35V 0-2A 2 Meters	£30
Thurlby PL320QMD	PSU 0-30V 0-2A Twice	£160 – £200
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HP33120A	Function Generator 100 microHz – 15MHz	£350
HP53131A	Universal Counter 3GHz Boxed unused	£600
HP53131A	Universal Counter 225MHz	£350
Audio Precision	SYS2712 Audio Analyser – in original box	POA
Datron 4708	Autocal Multifunction Standard	POA
Druck DPI 515	Pressure Calibrator/Controller	£400
Datron 1081	Autocal Standards Multimeter	POA
ENI 325LA	RF Power Amplifier 250kHz – 150MHz 25W 50dB	POA
Keithley 228	Voltage/Current Source	POA
Time 9818	DC Current & Voltage Calibrator	POA

HP/Agilent HP 34401A Digital
Multimeter 6½ Digit £325 – £375

Marconi 2955B Radio
Communications Test Set – £800

How's your soldering?

Techno Talk

Mark Nelson

Not bad? That's admirable, but have you considered taking it up as a competitive sport? Soldering (or 'welding' as many of those English-not-a-first-language eBay ads like to call it) is not yet recognised by the international Olympics committee, but that day will surely come and some folk already take the challenge very seriously. Even better, you can improve your prowess at home, so what's not to like? Read on, get some philosophical instruction and prepare to have your intelligence insulted, all at no cost to you.

In February, IPC (no, not the original publisher of this magazine, but the global trade association serving the printed board and electronics assembly industries, their customers and suppliers) hosted the world's first hand-soldering and re-working championship. Staged in San Diego, California, it was won, within the allotted time of 75 minutes, by Ryosuke Matsunami of Japan, gaining him a $1,000 prize and a perfect score of 634 points, achieved without a single error. Contestants represented Britain, China, France, Germany, India, Indonesia, Japan, South Korea, Vietnam and Thailand. All competitors were tested with a fully functional soldered assembly, from which they were required to remove six specific components, remove old solder and clean the area of removed components.

Do try this at home?

If you fancy an even tougher challenge, the 'SMD Challenge' might appeal. It's for seriously advanced builders only (find more at **http://bit.ly/pe-may19-tt1**). Be warned that this kit involves hand-soldering an 0201 SMD package, which is only slightly larger than a grain of sand. There is also a 'Misery Edition' that uses even smaller 0805 packages and a QFN20 chip, which is much harder to hand solder. Neither I nor the publisher of this magazine accept any liability for your subsequent sanity damage.

Free advice

Nothing lasts forever and there's no such thing as a free lunch. There you go: two truisms and nothing to pay for these wise words. But what else do you get for nothing? Well, there's Usenet, the first free online host for discussion forums, and I was just going to tell you when it finally bit the dust. But to my amazement, I see that it's still going and has done since 1980, predating public use of the Internet by nine years. But as Wikipedia concedes, Usenet has diminished in importance versus Internet forums, blogs and mailing lists.

Google Groups, operational since 2001, is also free to use. It not only hosts groups of its own, but also acts as gateway and unofficial archive of Usenet postings dating back to 1981. If you sign up for a Google Groups account, you can use this to read and post to Usenet groups. However, many groups that began at Google Groups have now migrated elsewhere.

The new host of choice was Yahoo Groups, established under that name in 2000. It became extremely popular but then lost its glitter after the platform started to act very erratically following concerted hacking attacks. Users' email addresses and passwords were stolen and resold on a grand scale, leading to a mass exodus of groups to pastures new. Many were re-established on the Groups.io platform, launched in 2014 to much acclaim.

Engage BS alert now

With all of the foregoing, commercial sponsorship was unobtrusive to the users but some newcomer group hosting organisations have opted for more aggressive advertising, such as Tapatalk, which entered the fray in 2017. Reading advertisements is probably a reasonable price to pay for free-to-use forums (or fora, if you prefer), if you don't mind having your intelligence insulted from time to time. I don't mind too much, but the ads are not very well curated or selected, meaning you will be offered electronic gizmos such as the SkylinkTV antenna that lets you watch all of your favourite TV channels 'for free'. It's nothing to do with Sky TV and, with over 1.5 million units sold worldwide, it: 'Offers a more reliable and technologically advanced antenna than almost any other on the market. That means more channels, movies and shows for free, without any subscription and in a completely legal way. It's developed with military technology too and its use is entirely legal since the television operators themselves distribute the signal completely free of charge. Obviously television operators do not like it at all, but they cannot do anything about it. They are required by law to maintain this signal. And now, with advanced antennas like SkyLink Antenna, you can finally capture it.'

Dodgy claims

The 'amazing' SkyLink Antenna costs $36 plus postage and import taxes (which cost a small fortune). You also need to be aware that it's an indoor antenna (we all know how fabulously well those work!) and it will not pick up US channels in Europe, despite the misleading claims. Independent reviews (**http://bit.ly/pe-may19-tt3**) are universally negative and of course viewing premium programmes without payment is legal neither Stateside nor here in Europe. To read their extravagant claims go to: **http://bit.ly/pe-may19-tt2** and read item 2. Or see the same product on Amazon for £9.69 (search for VicTsing Freeview antenna). Spoiler alert: 96 disappointed customers gave it negative reviews on Amazon UK.

Toilet humour

Now go to item 11 on the same web page and you'll find something far more useful: an electronic 'Glow Bowl' that will transform any toilet into a multi-hued nightlight, changing colour constantly. It's awesome, it's 'motion-activated' (!) and it lights up the most important part of late-night relief, with a soft glow that doesn't wake up anyone else. It even features a built-in air freshener that leaves your bathroom smelling great. Mmm, nice.

Credit where it's due: this time the advertisement is tailored to the British market and lets you pay just £15.62, using PayPal. But my sleuthing discovered a better deal on Amazon, where it costs (wait for it) £2.36 delivered free (search for Sumbay nightlight).

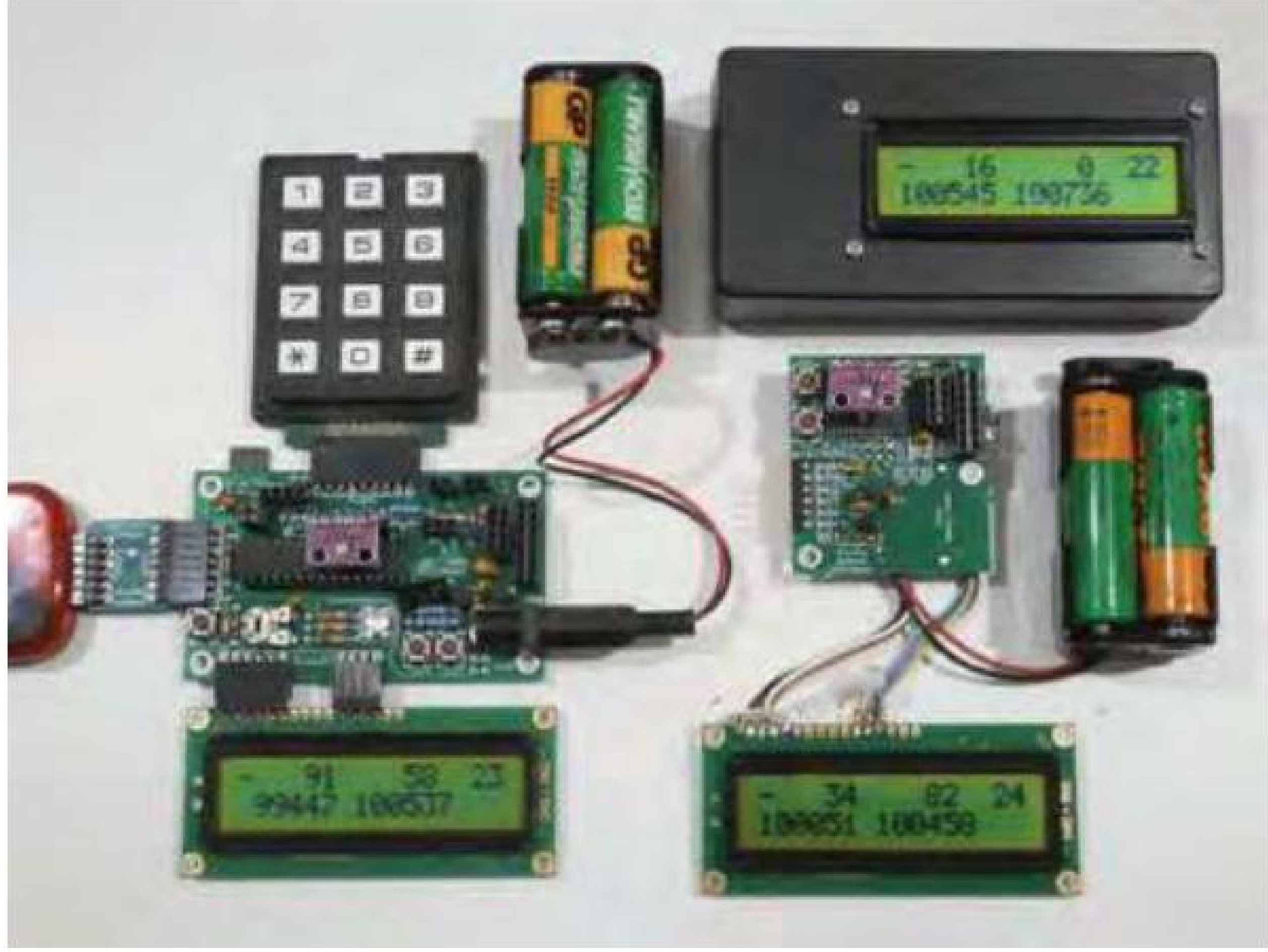
Compliance notice

Other brands of novelty gifts are available and products mentioned here are not endorsed by this magazine. 100% guaranteed not to contain nuts, probably.

Experimenting with BMP280

Temperature and Air Pressure Chip

by Peter Brunning

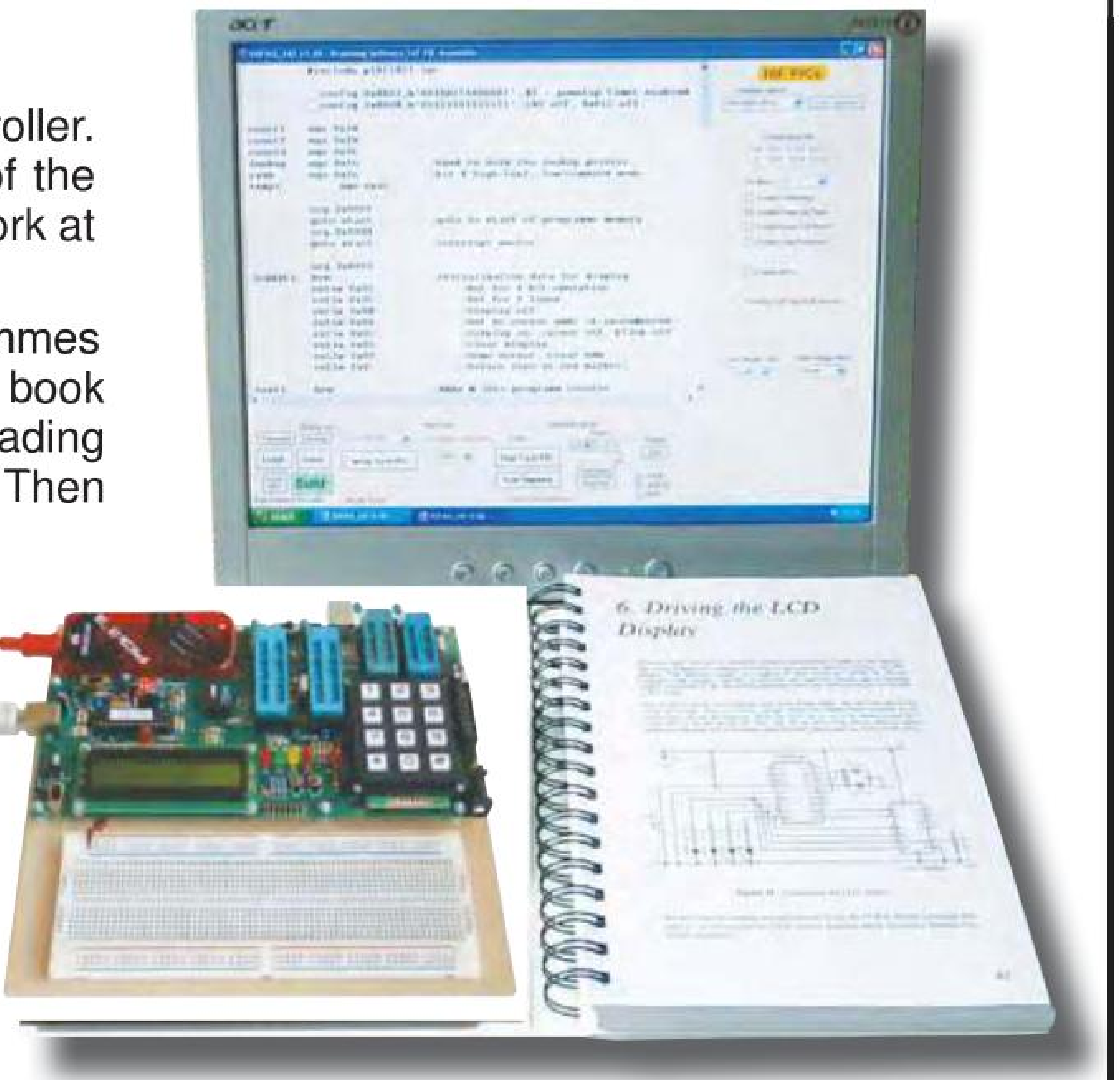


For this short course you need one of the GPIC28 training circuits shown in the picture, a PICkit 3, the 100-page book, and our BMP280 18F PIC C software library. Ideally, you will already have some experience, but the book does start with essential PIC C revision.

Then we study the BMP280 and learn to set up SPI serial communication. We experiment with code to display ID, raw temperature and calibration data. We use this data to calculate the temperature and then expand this to include air pressure. Air pressure reduces if we walk up a hill. We use this fact to calculate and display height gain.

Finally, we edit the code to display the total height gained when walking in hilly countryside. Both GPIC28 training circuits shown in the picture run the same code. The GPIC28sov4 on the right is supplied as a kit with the PIC already fitted. You need good soldering and workshop skills to fit this into a box as in the picture. The expected release date is early April 2019. Prices will be published as soon as possible. See our website for latest info.

The Brunning Software P955H PIC Training Course



We start by learning to use a relatively simple 8 bit PIC microcontroller. We make our connections directly to the input and output pins of the chip and have full control of the internal facilities of the chip. We work at the grass roots level.

The first book teaches absolute beginners to write PIC programmes using assembler which is the natural language of the PIC. The first book starts by assuming you know nothing about PICs but instead of wading into the theory we jump straight in with four easy experiments. Then having gained some experience we study the basic principles of PIC programming, learn about the 8 bit timer, how to drive the alphanumeric liquid crystal display, create a real time clock, experiment with the watchdog timer, sleep mode, beeps and music. Then there are two projects to work through. In the space of 24 experiments two project and 56 exercises we work through from absolute beginner to experienced engineer level using the latest 8 bit PICs (16F and 18F).

The second book introduces the C programming language for 8 bit PICs in very simple terms. The third book Experimenting with Serial Communications teaches Visual C# programming for the PC so that we can create PC programmes to control PIC circuits.

In the fourth book we learn to programme 32 bit MX PICs using fundamental C instructions. Flash the LEDs, study the 16 bit and 32 bit timers, write text to the LCD, and enter numbers using the keypad. This is all quite straightforward as most of the code is the same as already used with the 8 bit PICs. Then life gets more complex as we delve into serial communications with the final task being to create an audio oscilloscope with advanced triggering and adjustable scan rate.

The complete P955H training course is £259 including P955H training circuit, 4 books (240 × 170mm 1200 pages total), 6 PIC microcontrollers, PIC assembler and programme text on CD, 2 USB to PC leads, pack of components, and carriage to a UK address. (To programme 32 bit PICs you will need to plug on a PICkit3 which you need to buy from Microchip, Farnell or RS for £43).

Prices start from £175 for the P955H training circuit with books 1 and 2 (240 × 170mm 624 pages total), 2 PIC microcontrollers, PIC assembler and programme text on CD, USB to PC lead, and carriage to UK address. (PICkit3 not needed for this option). You can buy books 3 and 4, USB PIC, 32 bit PIC and components kit as required later. See website for details.

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Net Work

Alan Winstanley

Alan Winstanley says dodge the ‘Sunk Cost Fallacy’ of keeping filament bulbs and embrace the brave new world of LED lighting – plus, 5G and microsatellites are about to change our digital world.

The cost of electrical energy is a heavy burden for many households to shoulder, but any efficiency savings that consumers make are soon swallowed up by the escalating costs charged by power utilities. A recent discussion in our forum area on EEWeb (**www.eeweb.com/forum/category/epe-magazine**) highlighted the disparities in power prices levied in different countries; our friends in the US pay about $0.10 per kWh or half the cost that we Brits are forced to pay, while Australians and New Zealanders tell me their unit costs are roughly on parity with UK prices but they face far higher ‘standing charges’ every month.

Lamenting the lumen

In the hope of reducing electricity bills, the author has, like many readers, taken the plunge by replacing many old incandescent lights with modern LED bulbs. The lighting market faced turmoil when awful first-generation compact fluorescent tubes were the only alternatives available, but thankfully LED technology has come on in leaps and bounds, and buyers now enjoy more choice than ever. Some trial and error is to be expected as some promising-looking LED bulbs offer disappointing levels of output in practice, leaving buyers ruefully comparing ‘lumens’ against good old watts. A wide range of ‘LED filament’ bulbs are also sold that mimic gloomy vintage-style bulbs, but are probably best kept for ornamental use.

I probably should have bitten the bullet sooner. It can be hard to throw away a few dozen perfectly good, paid-for, incandescent light bulbs – especially traditional-quality ones that might even contain a Ballotini fuse, unlike the cheap end-of-an-era imports that I was now throwing away. Should one wring every last minute of working life out of a light bulb (due to the ‘Sunk Cost Fallacy’) until it blows, or write them off today and start benefiting from immediate and ongoing savings instead?

Behavioural economic theory explains that ‘sunk costs’ represent cash that has been spent and cannot be recovered under any circumstances. Hence the pain – investment in those treasured old filament bulbs can be considered as ‘sunk’ and gone for ever, and the only benefit in keeping them until they blow is to postpone the inevitable cost of replacement, whilst all the time costing ten times more to run than LEDs do.

False savings

Not wanting to throw away old lighting is a human, emotional and hence totally irrational thing. The same question taxes many British local authorities as they ponder converting high-pressure sodium street lights to those migraine-inducing LED bulbs that pierce our streets with their icy fingers of ‘cold’ light. Many enlightened councils have converted their street lights to LEDs, but some cash-strapped councils switch off street lights early to save money instead! For consumers, online auction sites are bursting with LED bulbs, some of them of very dubious quality, being unmarked and unbranded and perhaps with phony CE marks. They can have very poor heat dissipation properties or suffer from early failure due to penny-pinching with cheap components or poor design. For consistent performance it’s probably best to stick with reputable brands. A useful cost-saving calculator is provided online by Philips, for example, who manufacturer one of the world’s most comprehensive ranges of LED bulbs, see: **http://bit.ly/pe-may19-nw1**

Simply dial in the power, shape and base fitting and typical savings will be displayed. A huge range is sold by Lampshop Online, CPC, LEDbulbs.co.uk, BLT Direct and others. If you haven’t invested yet, then now is a good time to convert to LEDs and start making some immediate savings. Having said that, I noticed that one of my new long-life LED bulbs has failed after just a few months. So much for progress!

5G – go for throttle up

An old acquaintance lives in one of Britain’s largest cities and he had become alarmed by the growth of mobile phone masts springing up in his locality. He worried about the risks of electromagnetic radiation on his health, and had taken the extreme steps of lining his attic with chicken wire to protect the household from potentially harmful emissions generated, so he thought, by some nearby phone masts that menaced his property.

While most of us take wireless-based communication for granted, others have genuine concerns about the associated health risks. In the worst cases, some claim to suffer from electromagnetic hypersensitivity (EHS) and find life near overhead pylons, phone masts or electric cables unbearable. Unfortunately for them, the growth of radio traffic in an overcrowded radio spectrum continues unabated: we live in a cloud of electromagnetic smog, and life for EHS sufferers or my friend living in his home-made Faraday cage could be about to get worse.

In wireless communications the buzz word of the moment is ‘5G’ and the tech markets are queuing up to bring us the fifth generation of network hardware. 5G potentially operates at several gigabits per second with near-instantaneous latency values, reports the independent website 5GUK Ltd (**www.5g.co.uk**). Apart from enhancing mobile services the 5G industry claims it could also control IoT devices or even driverless cars, thanks to 5G’s high speed and super-responsive nature.

htc
18
TELSTRA
5G
HTC 5G Hub

HTC has pre-announced their 5G Mobile Hub, one of the first 5G consumer peripherals to hit the market.

Health risks from 5G?

5G's higher frequencies could, however, give rise to some as yet unknown health risks. A group of 180 EU scientists recently called for a moratorium on the rollout of 5G until its health implications had been independently assessed. The group cites the need for building many more 5G antennas in built-up areas as a major factor, due to the shorter distance that 5G waves can travel, and reminds us that only 5Gs's benefits have been praised by the media, but the consequences for the health of humans, plants and animals have not yet been discussed at all. One argument being made is that, as 5G masts generate lower power transmissions, a much denser mesh of them will therefore be required. Typically, the EU's response to the matter of electromagnetic radiation is that public safety is a matter for EU member states themselves to sort out, meaning there is no EU-wide approach to assessing the safety of 5G.

Similar concerns about 5G's health implications have been voiced in the US. In February, a Senate committee hearing on the future of 5G learned that no independent studies on the impact of 5G on human health had been undertaken at all, and US Senator Richard Blumenthal summed it up concisely: he believed Americans deserved to know what the health effects are, but the 5G industry was effectively 'flying blind' as far as any health or public safety assessments were concerned (see: **https://youtu.be/ekNC0J3xx1w**).

Coming to a hub near you

The technology sector industry is busily tantalising us with the potential of 5G, such as Vodafone's demo last year of the UK's first 5G holographic call (see: **https://youtu.be/Ilq2qtFHTf8**). It would be churlish not to applaud early efforts like these, which uses 3D video capture and VR to offer a young football fan a memorable moment or two with a holographic football star, but the lack of any existing infrastructure hasn't got in the way of manufacturers queuing up to advance-sell us the latest 5G kit. First off the starting blocks is Taiwan's HTC, which trailed its new rechargeable 5G Hub for home or business use. The Hub offers a 5-inch HD touch screen, 4K video streaming, connectivity to 20 devices and 'Cloud VR support in the future'. Plus, HTC reckons it will replace your home Wi-Fi router and all the unnecessary cables that go with it, being able to stream directly to a VR headset without needing to attach a PC. Looking at my motley collection of networked Internet radios, PCs, NAS drives, IoT gadgets, Google Minis, IP security cameras, phones and laptops, I'm not expecting to throw that lot away any day soon, but *PE* readers can expect to see a barrage of 5G hype heading their way well before the end of this year. Prices of the mobile hub have yet to be announced: HTC's UK pressroom has not been updated since 2017, but details were found on their US site. You can see what the HTC 5G Hub is about with a short Youtube video at: **https://youtu.be/SGeNk9FiJMI**

Already, the benefits of 5G are being lauded with plenty of vapourware but there is next to no working hardware on retail sale and no assessments of public health and safety have taken place either. 5G promises much, but who will be an early adopter?

OneWeb aims for the stars

There are many places on earth that are Internet 'not-spots' but the landscape could be about to change, thanks to space telecoms company OneWeb, which is set to bring high speed to the remotest areas of our planet. Headquartered in the UK, OneWeb has already started building a constellation of some 600 satellites, possibly scaling it up to 900 in the future, with the intention of serving remote users in homes, schools, businesses or even on the move. OneWeb Satellites (**http://onewebsatellites.com**) is a joint venture between Airbus Industries and OneWeb, and they are taking satellite production to an industrial level. It used to take hundreds of engineers months to hand-build a single space satellite, says OneWeb, but by using a modular satellite design they can mass produce a small 150kg satellite every eight hours – or 15 a week – at its factory in Florida that is currently being commissioned.

OneWeb's satellites will aim to provide low latency, high throughput and global coverage. They are said to be easy to manufacture, lighter in weight and cheaper to launch and use fewer components. The first six satellites in OneWeb's constellation were launched successfully in February aboard a single Soyuz rocket, and one can only marvel at the engineering that was behind the six-satellite separation sequence shown on Youtube at: **https://youtu.be/MnJZ3gJ9bAY**

Up to 36 satellites could be released per launch in the future. Helped by some UK Space Agency Funding, OneWeb aims to develop novel AI techniques to manage their fleet of satellites as well as ways of interfacing with terrestrial 5G communications networks, holding out the promise of OneWeb playing a key role in delivering a far-reaching 5G service down on the ground.

OneWeb successfully launches the first six of its fleet of 600 satellites in February.

Quantum leap

Flexibility and adaptability are keywords for a new wave of satellites that are set to revolutionise space-bound telecoms. Eutelsat's new Quantum satellite is a product of Airbus Industries and the UK and European Space Agencies, and has been dubbed a 'software-driven satellite', said to be the first satellite ever produced that can be completely reprogrammed in orbit to cover different (or multiple) geographical areas, as explained at: **https://youtu.be/0iKgOahW5y0**

Quantum can also be re-tasked to deal with governmental duties, such as public protection or disaster recovery, for example. The Quantum's platform has been built in Britain by Surrey Satellite Technology (SSTL) and the satellite is currently being readied by Airbus in Toulouse.

Maplin opens its doors

The former UK electronics mail-order house and retailer, Maplin, has finally re-opened for business under an entirely new business model. Now trading as Maplin Online Ltd, the new web-only presence is a valiant attempt to keep an old brand alive, but the firm is focussing on gadgets, IoT, makers and STEM, gaming, smart tech, AV and similar lines instead. Here at *PE* we believe that having more choice is a good thing for hobbyists and students, and we wish the new enterprise well in its endeavours. For all your component-level needs, however, please do continue to support our very hard-working advertisers who strive to offer readers a comprehensive mail-order service, often going the extra mile to provide constructors with a friendly helpful and informed source of kits and parts.

See you next month for more tech commentary from *Net Work!*

The author can be reached at:
alan@epemag.net

Look after your lithiums!

2× 12V Battery Balancer

By Nicholas Vinen

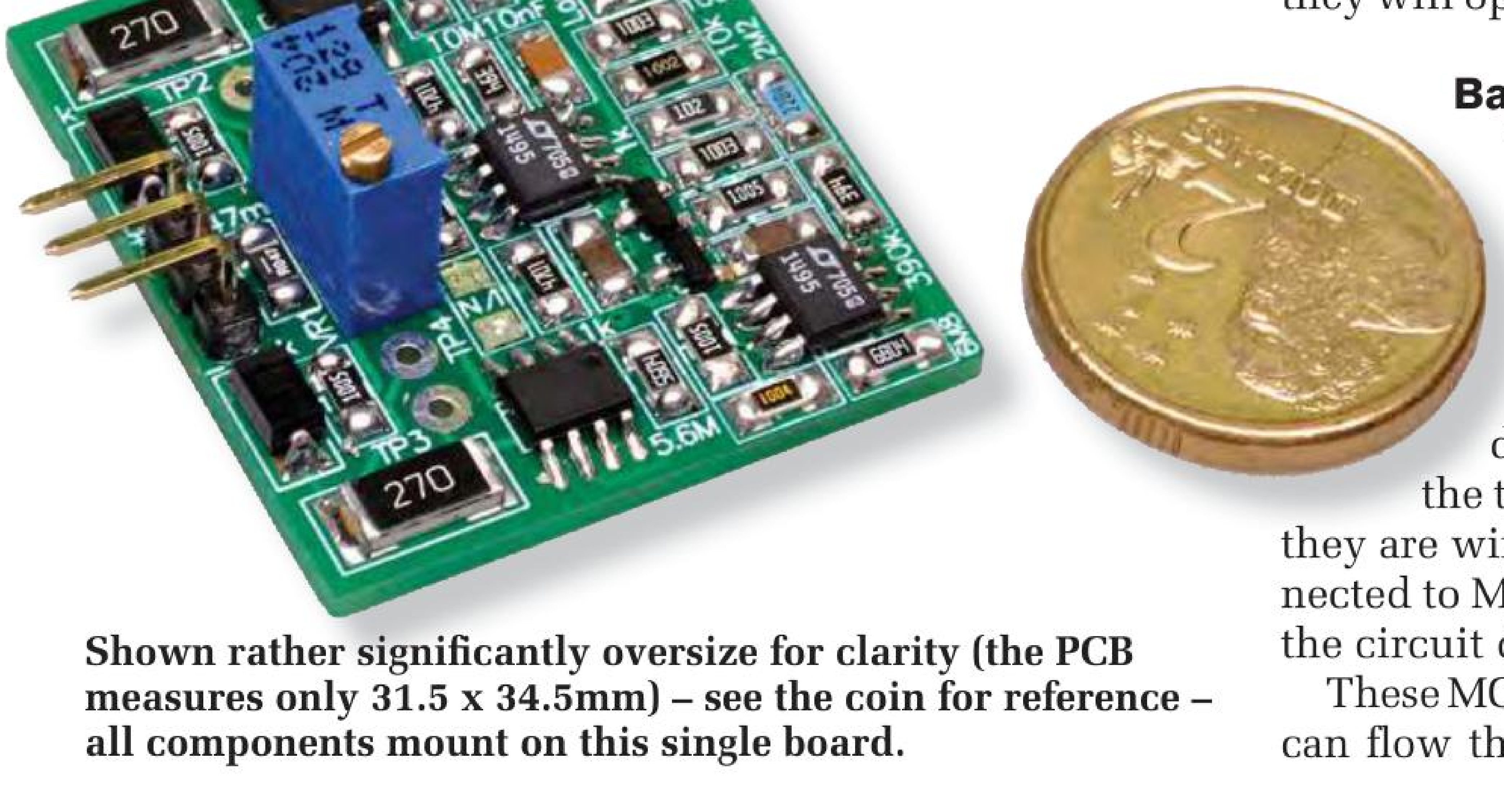
Two 12V batteries are often significantly cheaper than one equivalent 24V battery, but you need to be careful connecting batteries in series as their voltages and state-of-charge may not be identical. The difference in voltage can increase over time, leading to battery damage from over-charging and/or under-charging. This compact, low-cost device keeps them balanced so that they last a long time.

If you're in the market for a 24V battery then you have probably discovered that it costs a lot more than two 12V batteries. The obvious solution is two 12V batteries in series, but as our introduction explains, that's a solution that brings a new set of problems. Fortunately, an economical answer is within your reach – a 'battery balancer'.

Our version can't handle quite as much current as some commercial designs because it lacks a large heatsink, but you can easily parallel several of our balancers if you need a higher current capacity and the cost would still be quite reasonable.

It can be used with pretty much any battery chemistry, as long as the battery voltages will stay within the range of 5-16V.

Balancing is most critical with lithium-based rechargeable batteries because they tolerate overcharging much less than a similar lead-acid battery would.



Shown rather significantly oversize for clarity (the PCB measures only 31.5 x 34.5mm) – see the coin for reference – all components mount on this single board.

This design incorporates a low-voltage cut-out which prevents the batteries being discharged too far if it is unable to keep them balanced and its very low quiescent current of under 0.02mA means it will have virtually no effect on battery life.

It also incorporates an LED to show when it is monitoring the battery voltages and two more LEDs to show when one or the other is being discharged or shunted.

By default, the low-voltage cut-out is set up so that the batteries are only balanced when they are being charged. However, there are definitely situations where you might also want the batteries to be balanced during discharge.

In that case, you just need to change a resistor or two in order to adjust the cut-out threshold so it is near the minimum battery voltage. In this case, the cut-out will still act to protect the batteries but will allow balancing during charging and discharge, right down to that lower threshold.

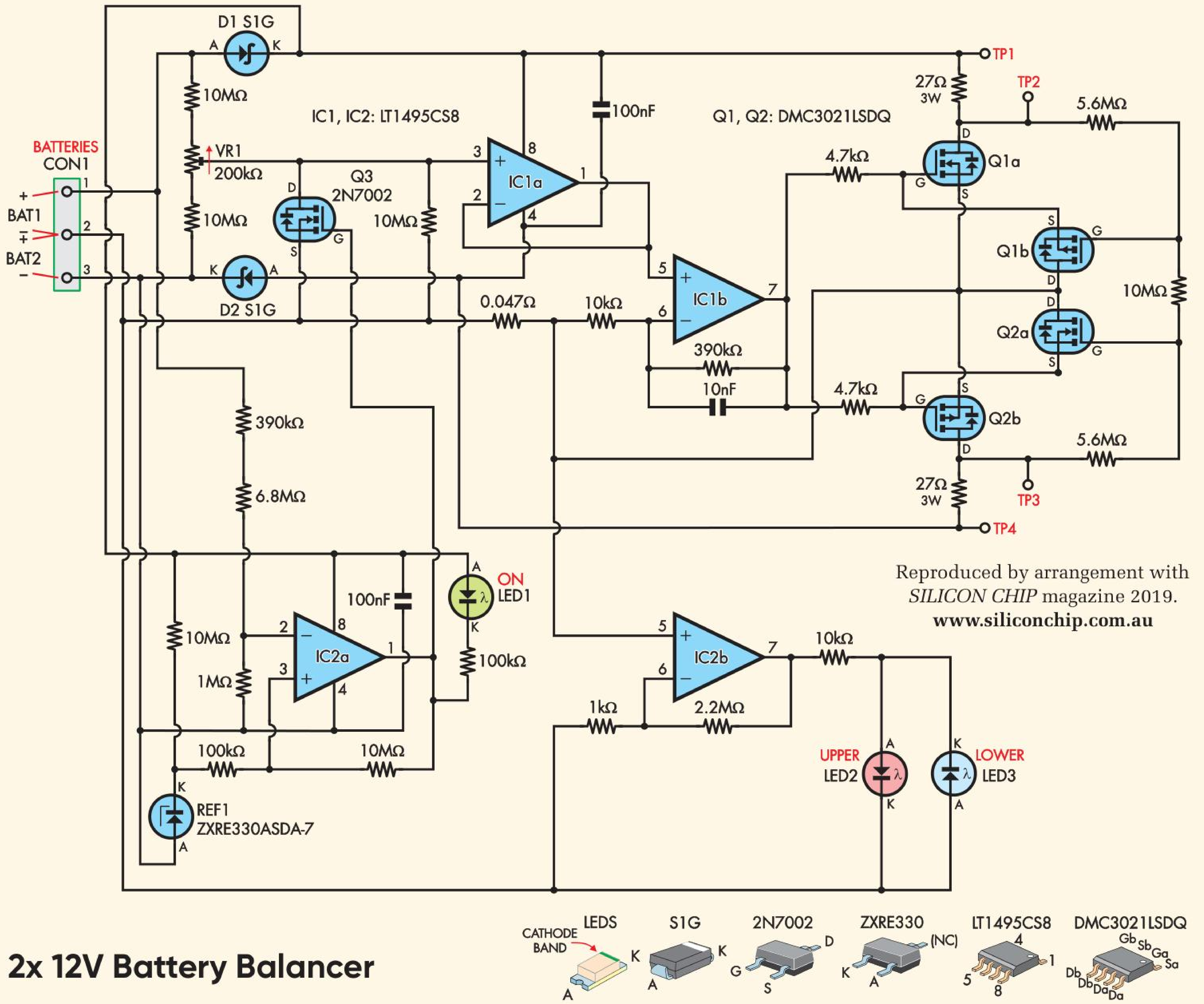
It's a compact unit at just 31.5 × 34.5 × 13mm, so you can tuck it away inside just about any device. And if the 300mA balancing current is not sufficient for your purposes, all you need to do is wire two or more units in parallel and they will operate in concert to keep the batteries balanced.

Balancing operation

There are two sections to the circuit; the balancer and the low-battery cut-out. The entire circuit is shown in Fig.1, with the balancing circuitry in the top half and the low-voltage cut-out below.

Starting with the balancing section, schottky diodes D1 and D2 are connected in series with the two batteries so that no damage should occur if they are wired up incorrectly. These diodes are then connected to MOSFETs Q1a and Q2b at the right-hand side of the circuit diagram, via a pair of 27Ω 3W resistors.

These MOSFETs are normally switched off and no current can flow through them. If the voltage across one battery


D1 S1G
A
K
TP1
27Ω
3W
TP2
5.6MΩ
10MΩ
100nF
IC1, IC2: LT1495CS8
Q1, Q2: DMC3021LSDQ
BATTERIES
CON1
VR1
200kΩ
3
8
IC1a
1
2
4
4.7kΩ
Q1a
D
G
S
Q3
2N7002
D
S
G
10MΩ
BAT1
BAT2
1
2
3
10MΩ
Q1b
K
A
D2 S1G
5
IC1b
7
6
0.047Ω
10kΩ
390kΩ
10nF
4.7kΩ
Q2a
Q2b
10MΩ
5.6MΩ
27Ω
3W
TP3
TP4
390kΩ
6.8MΩ
100nF
ON
LED1
A
K
100kΩ
10MΩ
2
8
IC2a
1
3
4
1MΩ
5
IC2b
7
6
10kΩ
1kΩ
2.2MΩ
UPPER
LED2
LOWER
LED3
100kΩ
10MΩ
REF1
ZXRE330ASDA-7
LEDS
CATHODE BAND
S1G
2N7002
ZXRE330
(NC)
LT1495CS8
DMC3021LSDQ
Gb Sb Ga Sa
Db Db Da Da

2x 12V Battery Balancer

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Fig.1: the circuit for the *Battery Balancer*, shows the balancing section at top and low-voltage cut-out at bottom, based around dual micropower op amps IC1 and IC2 respectively. IC1 drives dual MOSFETs Q1 and Q2 to perform balancing when necessary; IC2 drives the indicator LEDs and disables IC1 using MOSFET Q3 when the battery voltage is low.

rises by more than 100mV above the other, the MOSFET across that battery is switched on.

If the battery is being charged, this has the effect of shunting some of the charge current around that battery so that it receives a lower charging current than the other, decreasing the voltage differential over time, as the battery with the lower voltage is then receiving more charging current.

If the unit is operating while the battery is not being charged, the effect is to slightly discharge the battery with the higher voltage until they are closer in voltage.

It's a linear circuit, so the shunt current is proportional to the difference in voltage. As the imbalance rises, so does the shunt current until the limit of around 300mA is reached. This is to prevent the MOSFET and resistor from overheating.

Detecting a voltage difference

A resistive divider comprising two 10MΩ resistors and a 200kΩ trimpot VR1 is connected across the battery, before diodes D1 and D2 so that their forward voltage does not affect the calculation of the difference in voltages.

VR1 is adjusted so that the voltage at its wiper is exactly half that of the total battery. This half-battery voltage is buffered by voltage-follower op amp IC1a. This op amp has a very high input resistance of around 40GΩ, resulting in a low input bias current of approximately 250pA, so the high values of these resistors (chosen to minimise the quiescent current) will not result in a large error voltage. The other half of the dual op amp (IC1b) compares the voltage at the junction of the two batteries (from pin 2 of CON1) to the output voltage from IC1a. If the upper battery has a higher voltage than the lower battery then the half-battery voltage will be higher than the voltage at pin 2 of CON1. That means that the voltage at non-inverting input pin 5 will be higher than at the inverting input, pin 6.

As a result, IC1b's output will swing positive. The ratio of the 390kΩ feedback resistor to the 10kΩ resistor that goes to the battery junction (ie, 39:1) means that the output will increase by 40mV for each 1mV difference in battery voltages.

Once the voltage at output pin 7 has risen by a couple of volts, N-channel MOSFET Q1a will switch on because its gate is being driven above its source, which connects to pin 2 of CON1 via a low-value shunt resistor (47mΩ).

So current will flow from the positive terminal of the upper battery, through diode D1, the 27Ω 3W resistor, MOSFET Q1a and then the 47mΩ resistor to the negative terminal of the upper battery.

Once this current starts to flow, it will also develop a voltage across the 47mΩ resistor, which will increase the voltage at pin 6 of IC1b, providing negative feedback. This feedback is around 1mV/20mA, due to the shunt value.

This prevents Q1a from switching fully on. Rather, its gate voltage will increase until the current through the 47mΩ resistor cancels out the difference in the two voltages.

Features and specifications

Minimum battery voltage	*5V*
Nominal battery voltage	*12-13V*
Maximum battery voltage (fully charged)	*16V*
Battery voltage difference for balancing to start	*approximately 100mV*
Battery voltage difference for max balancing current	*approximately 130mV*
Maximum balancing current	*approximately 300mA (multiple units can be paralleled)*
Maximum balancing power	*approximately 4.5W (multiple units can be paralleled)*
Maximum recommended charging current	*10A per unit*
Quiescent current	*< 20µA*
Low-voltage cut-out threshold	*27V (can be changed)*
Low-voltage cut-out hysteresis	*0.25V*

Hence, the maximum shunt current of 300mA will be achieved with an imbalance around 130mV (100mV + 300mA × 0.047Ω × 2).

The 10MΩ resistor between pin 3 of IC1a and pin 2 of CON1 serves mainly to prevent the balancer from operating should the junction of the batteries become disconnected from CON1. It also makes setting the unit up and adjusting VR1 easier. It has a negligible effect on the voltage at pin 3 since there's normally such a small voltage across it.

Current limiting

Had we specified 68Ω resistors in series with Q1a and Q2b (rather than 27Ω), there would be no need for additional current-limiting circuitry since the resistors would naturally limit the balancing current within their dissipation ratings.

However, this would mean that at the maximum balancing current, all the dissipation would be in this resistor and none in the MOSFET, meaning the maximum current would be 200mA [14V ÷ 68Ω].

We realised we could increase this by 50% by splitting the dissipation between the MOSFET and its series resistor. The resistor has a 3W rating while the MOSFET has a 2W rating, giving the possibility of a total of just under 5W. With a battery voltage of 29V and a balancing current of 300mA, dissipation is around 2.7W in the resistor and 1.7W in the MOSFET.

We achieve this dissipation sharing by preventing the MOSFET from turning on fully and using a lower value limiting resistor. This is the purpose of Q1b and the three resistors between TP2 and TP3.

These resistors bias the gate of Q1b at a voltage that's initially about halfway between the negative and positive terminals of the upper battery (ie, at a voltage between that of pins 1 and 2 of CON1). However, as the balancing current for the upper battery increases, the voltage at the junction of the 27Ω resistor and Q1a drops, and therefore, so does Q1b's gate voltage.

Q1b is a P-channel MOSFET, and so it switches on when its gate is a few volts below its source terminal. The source terminal is connected to the gate of Q1a, which is about 2V above pin 2 of CON1 when Q1a is in conduction.

So as the current through Q1a builds and Q1b's gate voltage drops, eventually Q1b begins to conduct, pulling the gate of Q1a negative and cutting it off. This forms a negative feedback path and due to the gate capacitances, the circuit stabilises at a particular current level.

With 300mA through the 27Ω resistor, the voltage across it will be 8.1V [0.3A x 27Ω] and this translates to a gate-source voltage for Q1b of around –2V; ie, just enough for it to conduct current. The 4.7kΩ resistor between output pin 7 of IC1b and the gate of Q1a prevents Q1b from 'fighting' the output of the op amp too much.

Note that 8.1V is slightly more than half the typical voltage of one 12V battery and this is why the resistor dissipates slightly more than the MOSFET, in line with their ratings.

Balancing the other battery

The other half of the balancing is a mirror-image; for balancing the lower battery, MOSFET Q2b is a P-channel type and thus switches on when its gate is driven below its source. As with Q1a, its source is connected to the junction of the two batteries via the 47mΩ resistor.

When the lower battery voltage is higher than the upper battery, output pin 7 of IC1b goes negative, switching Q2b on.

The same current-limiting circuitry is present, but this time, Q2a is an N-channel MOSFET, so that as current builds through the lower 27Ω resistor and the voltage at the junction of it and Q2b rises, Q2a switches on and limits the current to a similar 300mA value, with roughly the same dissipation split between the two components.

A 10nF capacitor across IC1b's 390kΩ feedback resistor slows down its action so that it doesn't react to any noise or EMI which may be present at the battery terminals (eg, due to a switch-mode load).

It also prevents the circuit from oscillating due to the negative feedback and the action of the current limiters.

Under-voltage cut-out

Commercial battery balancers tend to only operate when the battery voltage is near maximum, as this is when they are being charged. That avoids the possibility of the balancer discharging the batteries when they are under load.

However, we're not convinced this is a good idea. It's possible to have a sufficient initial imbalance that one battery could be over-charged before the balancer even activates. And full-time balancing also has the advantage that it can start re-balancing the cells as soon as an imbalance occurs, which also avoids over-discharge and gives it more time for re-balancing.

There is one other advantage to having a higher under-voltage lockout threshold – it will prevent the balancer being triggered due to differing internal resistance of the batteries when under heavy load. This could create a voltage difference between the batteries even when they are at an equal state of charge.

If you want the balancer to be active, even when the batteries are not being charged, you still need the under-voltage lockout circuitry to prevent the balancer from over-discharging either battery. But in that case, you would change its threshold to be close to the fully discharged voltage of your combined battery.

For a pair of lithium-based 12V rechargeable batteries, this would normally be around 20V total. That's to protect the case where one battery has a failure (eg, shorted cell) which causes its voltage to drop dramatically. The under-voltage detection circuitry will then prevent the balancer from over-discharging the other battery in response, and potentially destroying it. See the section below on how to change the cut-out threshold if you want to take this approach.

The increased battery drain of the low-voltage cut-out section is only about 10µA. As a bonus, it drives the three LEDs to indicate when the balancer is operating and which battery is being shunted.

This is implemented using IC2a, another LT1495 op amp. Its positive supply is the same as for IC1a, but its negative

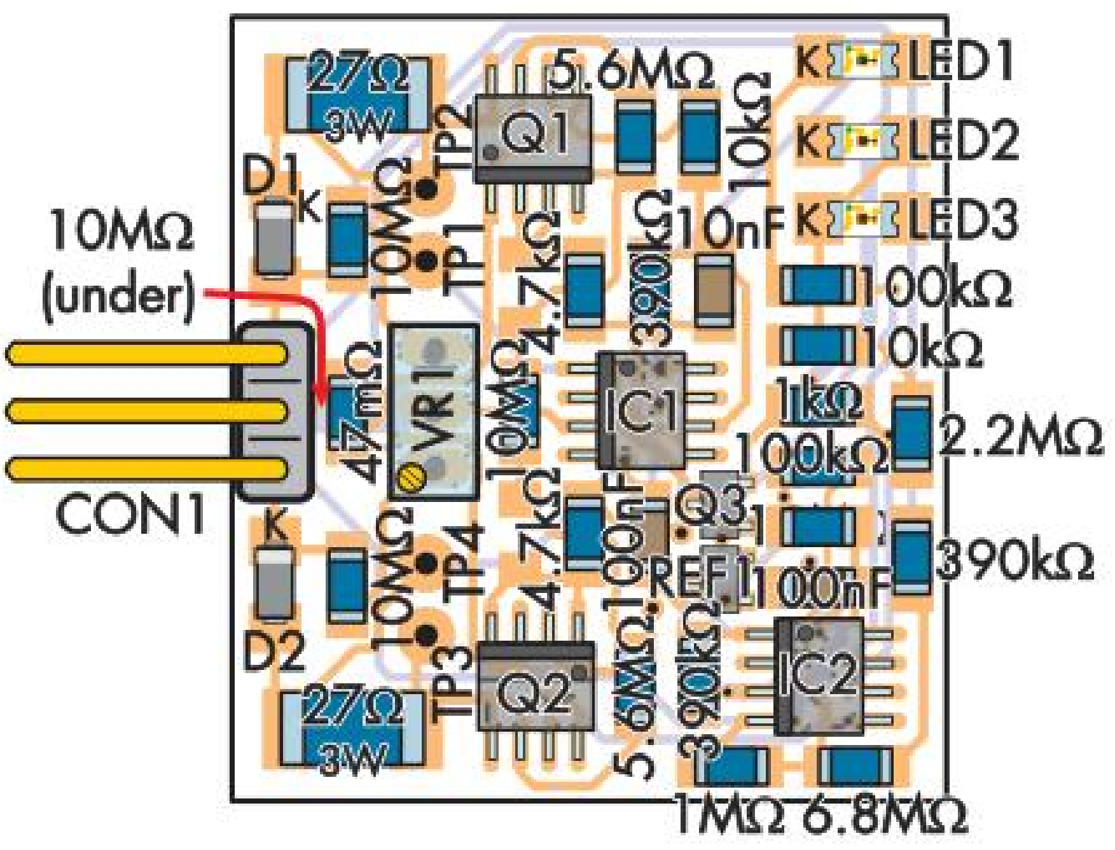
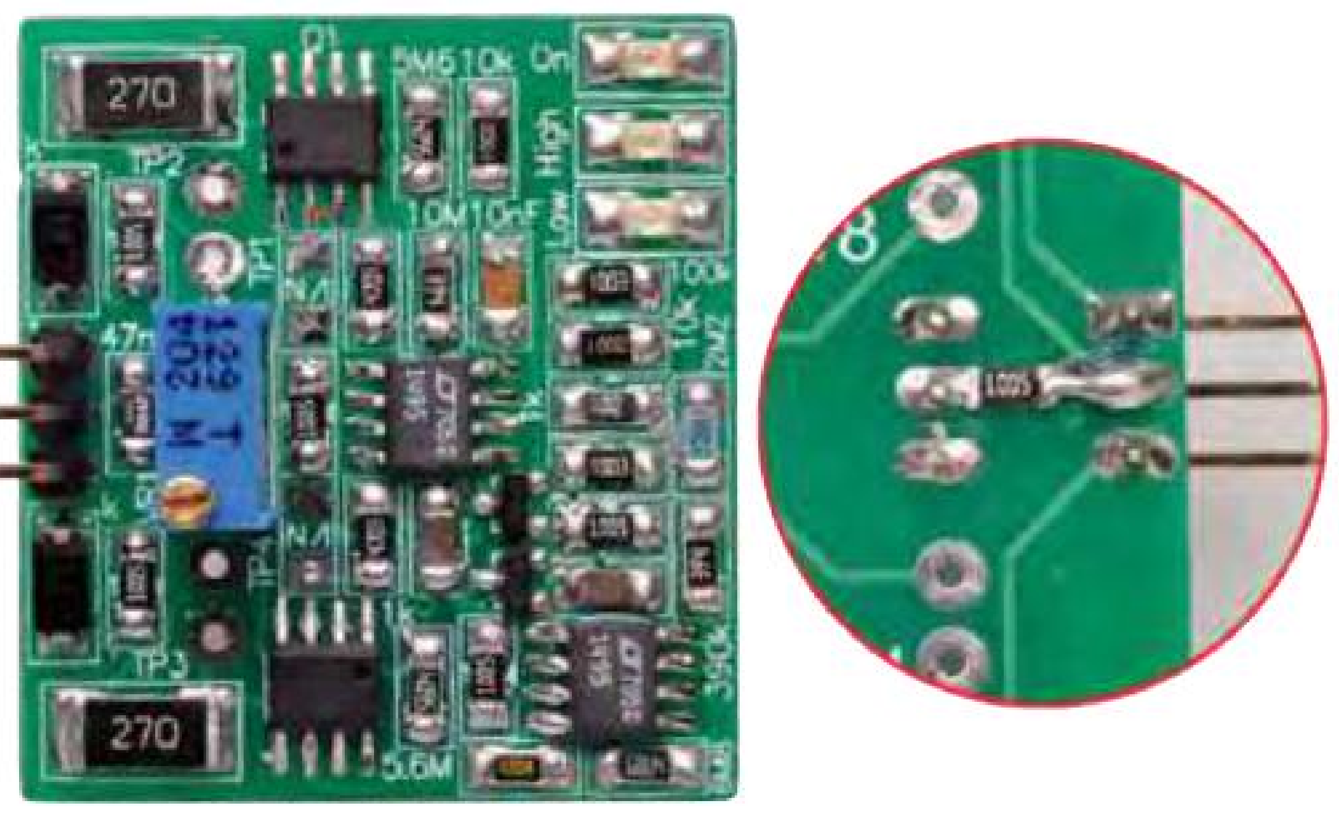


Fig.2: use the PCB overlay diagram at left and matching photo at right as a guide to assembling the PCB. Only one SMD component (a 10MΩ resistor) is soldered to the bottom, the rest go on the top as shown. The main aspects to pay attention to during constructon are that the semiconductors are correctly oriented and that you fit the resistors and capacitors in the correct locations.



supply is connected directly to the negative terminal of the bottom battery, allowing it to sense the total battery voltage (ie, between pins 1 and 3 of CON1) more easily.

This is done using a string of three resistors (390kΩ, 6.8MΩ and 1MΩ) connected across the batteries. These form a divider with a ratio of 8.19 [(390kΩ + 6.8MΩ) ÷ 1MΩ + 1]. The divided voltage from the battery is applied to inverting input pin 2 of IC2a.

A 3.3V reference voltage is applied to the non-inverting input at pin 3. This is provided by micropower shunt reference REF1, which is supplied with around 2µA via a 10MΩ resistor. The voltage at pin 2 of IC2a is therefore above the voltage at pin 3 when the battery voltage is above 27V [3.3V x 8.19].

When this is the case, output pin 1 of IC2a is driven low, pulling the gate of N-channel MOSFET Q3 to the same voltage as the negative terminal of the bottom battery. Since the source of Q3 is connected to the junction of the two batteries, Q3 is off and so does not interfere with the operation of the balancer.

However, should the total battery voltage drop below 27V, the output of IC2a goes high, switching on Q3 and effectively shorting input pin 3 of IC1a to the junction of the two batteries. This means that the voltages at pins 5 and 6 of IC1b will be equal (with no current flow through the 47mΩ resistor, as will quickly be the case), therefore preventing any balancing from occurring.

When the output of IC2a goes high, this also causes a slight increase in the voltage at its pin 3 input, due to the 10MΩ feedback resistor. This provides around 1% or 250mV hysteresis, preventing the unit from toggling on and off rapidly. In other words, the battery voltage must increase to 27.25V to switch the balancer back on.

When the output of IC2a is low and the balancer is active, IC2a also sinks around 0.25mA through LED1 and its 100kΩ series resistor, lighting it up and indicating the balancer is operating.

And when one or the other battery is being shunted, IC2b amplifies the voltage across the 47mΩ shunt by a factor of 2200 times. So if there is at least 20mA being shunted, that results in around 1mV across the 47mΩ resistor, which translates to 2.2V at output pin 7 of IC2b, enough to light up either LED2 or LED3. LED2 is lit if it's the upper battery being shunted, and LED3 if it's the lower battery.

Changing the cut-out voltage

To change the cut-out voltage, simply change the values of the 6.8MΩ and 390kΩ resistors using the following procedure. First, take the desired cut-out voltage and divide by 3.3V. Say you want to make it 24V. 24V ÷ 3.3V = 7.27. Then subtract one. This is the desired total value, in megohms. So in this case, 6.27MΩ.

This can be approximated a number of ways using standard values. For example, 3.3MΩ + 3.0MΩ = 6.3MΩ which is very close. So use these values in place of the 6.8MΩ and 390kΩ resistors.

Keep in mind there will still be around 1% hysteresis, so the switch-on voltage will be about 24.24V.

Two more examples would be a 22V cut-out, which would require 5.67MΩ total; you could use 5.6MΩ + 68kΩ. Or for a 20V cut-out, you would need 5.06MΩ which could be formed using 4.7MΩ + 360kΩ.

Paralleling multiple boards

As stated, one board can handle around 300mA and will dissipate up to around 4.5W. If you're using a 3A charger, that means it can handle a ~10% imbalance in charge between batteries (which would be unusually high).

However, with a 10A charger, it will only handle a ~3% imbalance, with a 20A charger ~1.5%, and so on. A greater imbalance could potentially lead to over-charging as the balancer can't 'keep up'. So if your charger can deliver more than 5A, you may want to consider paralleling multiple balancers and we would strongly recommend it for a charger capable of 10A or more.

When properly adjusted, the balancers will share the load. Realistically, one of them will start balancing first, but if it's unable to keep the imbalance voltage low, the others will quickly kick in and shunt additional current.

Since the only external connections are via 3-way pin header CON1, you could simply stack the boards by running thick (1mm) tinned copper wire through these pads and soldering them to each board in turn. You can then solder the battery wires to these wires.

Sourcing the parts

The PCB is available from the *Practical Electronics PCB Service*, board code 14106181.

All the other parts are available from Digi-Key. You can find the semiconductors on their website by searching for their part number and then narrowing down the list (eg, ignoring listings which are out of stock or only sold in large quantities). Visit: **www.digikey.co.uk**

For the other, more generic parts like SMD resistors, you can find them by searching for (for example) 'SMD resistor 1206 4.7k 1%' and then sorting the result by price. The cheapest part which matches the specifications should do the job just fine. But be careful – sometimes the search results include parts with different properties than you are expecting. You will need to skip over those.

Mouser, another large electronics retailer, will almost certainly have all the required parts too – **www.mouser.co.uk**

If either of these suppliers are out of stock for a device then it's also worth checking element14 (formerly Farnell) – see: **https://uk.farnell.com**

Construction

The *12V Battery Balancer* is built on a small double-sided PCB measuring 31.5 x 34.5mm and uses mostly surface-mounted parts. These are all relatively large and easy to solder. Refer to the overlay diagram, Fig.2, to see where each component goes on the board. Some of them (the ICs, MOSFETs, diodes and trimpot) are polarised, so be sure to fit them with the orientation shown.

There are two small SOT-23 package devices, MOSFET Q3 and voltage reference REF1. Fit these first. They look almost identical so don't get them mixed up; only the tiny coded markings on the top of each set them apart.

Tack solder the central pin to the pad in each case, then check that the other two pins are centred on their pads and that all pins are in contact with the PCB surface. If not, reheat the initial solder joint and nudge the part into place. Then solder the two remaining pins and add a little extra solder to the first pin (or a bit of flux paste and heat it) to ensure the fillet is good.

Next, solder IC1, IC2, Q1 and Q2. They are all in eight-pin packages and must be oriented correctly. Identifying pin 1 can be a bit tricky. For IC1 and IC2, you have to find the chamfered edge, which is quite subtle. Pin 1 is on that side. Q1 and Q2 have pin 1 marked by a much more clear divot in the corner of the package. But you can also orient IC1 and IC2 by matching the position of the markings up to our photo.

In each case, make sure the device is positioned correctly and tack solder one pin, then as before, check the locations of the other pads are correct and solder them before refreshing the first joint. If you accidentally bridge two pins with solder, use a little flux paste and some solder wick to clean it up.

The only remaining SMD parts which are polarised are diodes D1 and D2. Fit these now, ensuring the striped end goes towards the top edge of the PCB, as shown in Fig.2 and marked with 'K' on the PCB. Then solder the two 3W resistors in place. Follow with the remaining SMD ceramic capacitors and chip resistors, as shown in the overlay diagram.

For the two-pin devices, make sure that you apply the soldering iron long enough so that the solder adheres to the PCB and the component. Adding a little flux paste to the PCB pads before positioning the part will make this easier.

There is a single component on the underside of the board, a 10MΩ resistor positioned between CON1 and VR1. Solder it in place, but use a minimal amount of solder, so that you don't plug the through-holes underneath. You can add more solder later after CON1 and VR1 are in place.

All that's left is to solder trimpot VR1 with the adjustment screw oriented as shown, and a pin header for CON1. We used a normal pin header, but a polarised header would be a good idea if you're going to use a plug to make connection to the batteries so that it can't be accidentally reversed. If it is reversed, D1 and D2 should prevent damage but the balancer won't work!

Or you can solder the battery wires directly to these three pads. They only need to be rated to handle 300mA per board; medium-duty hookup wire should be more than sufficient, even if paralleling multiple boards.

Testing and set-up

Connect your batteries in series, then connect the negative-most terminal directly to the negative terminal on CON1. Do not connect the junction of the two batteries to the *Balancer* just yet.

Ensure that the total battery voltage is well above the threshold and that they are reasonably close to being balanced. You can ensure they are balanced by charging both independently and then connecting them in parallel via a low-value, high-power resistor (eg, 1Ω 5W) and leaving them for a few hours. The voltage across the resistor should drop to a very low level once their voltages equalise.

Now connect the most positive terminal to the positive pin of CON1 via a 1kΩ resistor and check that LED1 lights up. LEDs 2 and 3 should remain off. Measure the voltage across the 1kΩ resistor. It should be under 20mV. If it's under 5mV or over 20mV, disconnect the battery and check for errors in your PCB assembly or battery wiring.

Parts list
2× 12V Battery Balancer

1 double-sided PCB, coded 14106181, 31.5 x 34.5mm
3 3-way right-angle or vertical pin header (CON1)

Semiconductors
2 LT1495CS8 dual micropower op amps, SOIC-8 (IC1,IC2)
1 ZXRE330ASA-7 micropower 3.3V reference, SOT-23 (REF1)
2 DMC3021LSDQ dual N-channel/P-channel power MOSFETs, SOIC-8 (Q1,Q2)
1 2N7000 N-channel signal MOSFET, SOT-23 (Q3)
1 green LED, SMD 3216/1206 (LED1)
1 red LED, SMD 3216/1206 (LED2)
1 blue LED, SMD 3216/1206 (LED3)
2 S1G 1A schottky diodes or similar, DO-214AC (D1,D2)

Capacitors (all SMD 3216/1206 X7R ceramic)
2 100nF 50V (*measure value before installing!*)
1 10nF 50V (*measure value before installing!*)

Resistors (all SMD 3216/1206 1%)
6 10MΩ (Code 1005)
1 6.8MΩ (Code 6804)
2 5.6MΩ (Code 5604)
1 2.2MΩ (Code 2204)
1 1MΩ (Code 1004)
2 390kΩ (Code 3903)
2 100kΩ (Code 1003)
2 10kΩ (Code 1002)
2 4.7kΩ (Code 4701)
1 1kΩ (Code 1001)
2 27Ω 3W (SMD 6331/2512)
[eg, TE Connectivity 352227RFT]
1 47mΩ [eg, Panasonic ERJ-L08KF47MV]
1 200kΩ 25-turn vertical trimpot (VR1)

For tips and tricks on soldering SMD components, refer to the July 2010 *EPE* article: *How to Solder Surface Mount Devices*.

Assuming the voltage is within the specified range, remove or short out the 1kΩ test resistor and then connect the junction of the two batteries to pin 2 of CON1. LED2 and LED3 may light up. If so, rotate the adjustment screw in VR1 until they are both off.

Now check that there is no balance current flowing by measuring the voltage between TP1 and TP2, and between TP3 and TP4. In each case, the reading should be zero. If you get a non-zero reading between TP1 and TP2, current is flowing through Q1a. And if there's a voltage between TP3 and TP4, current is flowing through Q2b.

Since you started out with balanced voltages, this should not be the case, so adjust VR1 further until you get a zero reading across both pairs of test points. Ideally, VR1 should be adjusted to halfway between the point where the voltage starts to rise between one pair of test points, and the point at which the voltage rises across the other pair of test points. This ensures the balancing will be, for lack of a better word, 'balanced'!

The maximum reading you should get between one pair of test points should be 8.8V. Any more than that and you risk the resistor dissipation rating being exceeded. In this case, disconnect the batteries and change the 10MΩ resistor right next to VR1 on the top side of the board with a slightly lower value (eg, 9.1MΩ or 8.2MΩ) to reduce the current limit.

If that doesn't fix it then it's likely that the current-limiting circuitry is not working so you should check for soldering problems or faulty components.

Win an AVR-IoT WG Development Board

Practical Electronics is offering its readers the chance to win an AVR-IoT WG Development Board (AC164160), which makes it easy to deploy IoT devices to Google Cloud IoT Core's artificial intelligence and machine learning infrastructure.

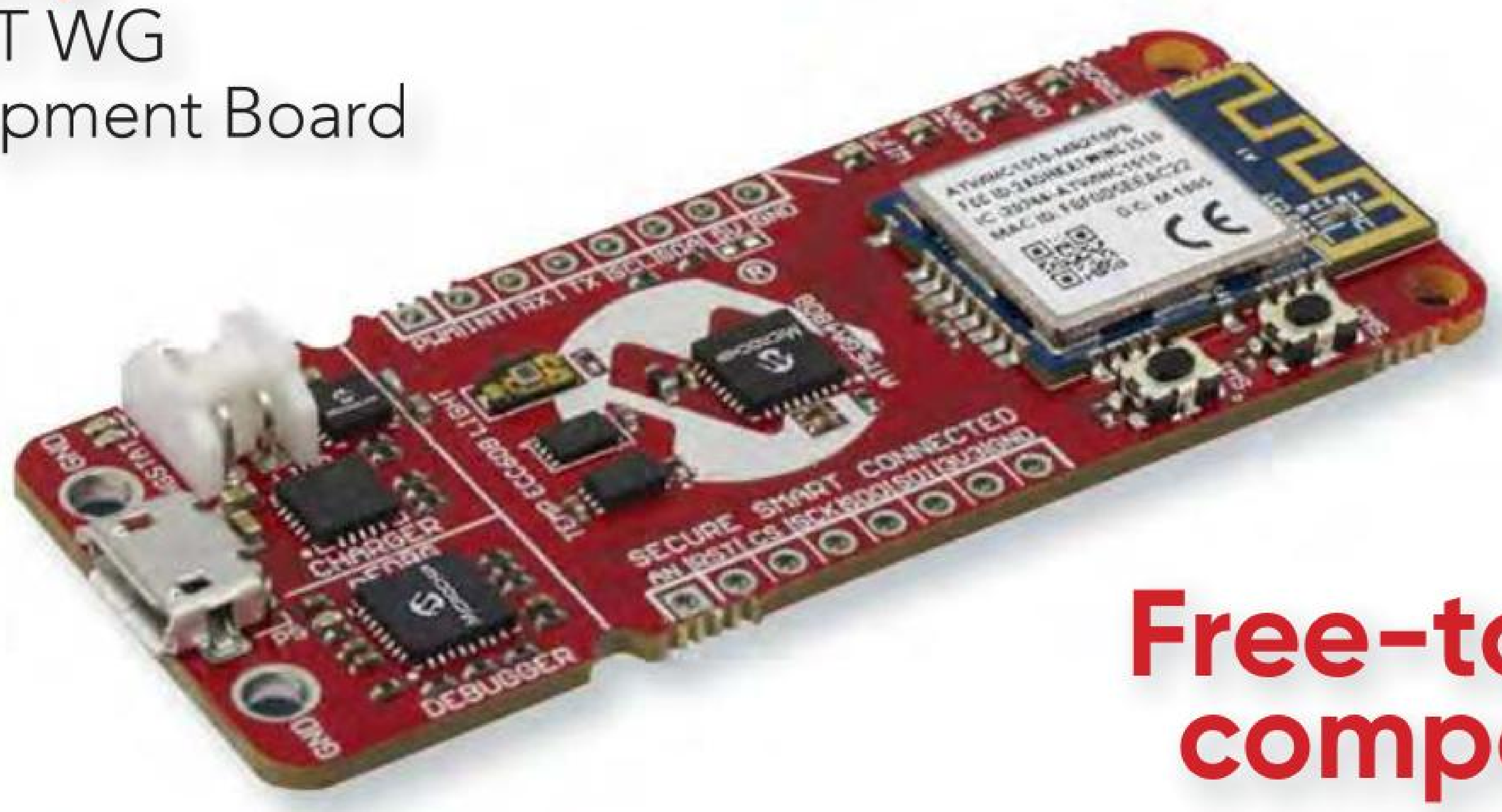
The AVR-IoT WG Development Board (AC164160) combines a powerful AVR microcontroller (the 8-bit ATmega4808 MCU), an ATECC608A CryptoAuthentication secure element IC and the fully certified ATWINC1510 Wi-Fi network controller to provide a simple and effective way to connect your embedded applications.

The AVR-IoT WG Development Board gives developers the ability to add Google Cloud connectivity to new and existing projects with a single click using a free online portal at www.AVR-IoT.com. Once connected, developers can use Microchip's rapid development tools, MPLAB Code Configurator (MCC) and Atmel START to develop and debug in the cloud.

The board combines smart, connected and secure devices to enable designers to quickly connect IoT designs to the cloud. The board is compatible with more than 450 MikroElektronika Click boards that expand sensor and actuator options.

Worth $29.00 (approx £22.30) each

Microchip
AVR-IoT WG
Development Board



Free-to-enter competition

How to enter

For your chance to win the AVR-IoT WG Development Board (AC164160), just visit:

http://page.microchip.com/EPE-AVR-IoT.html

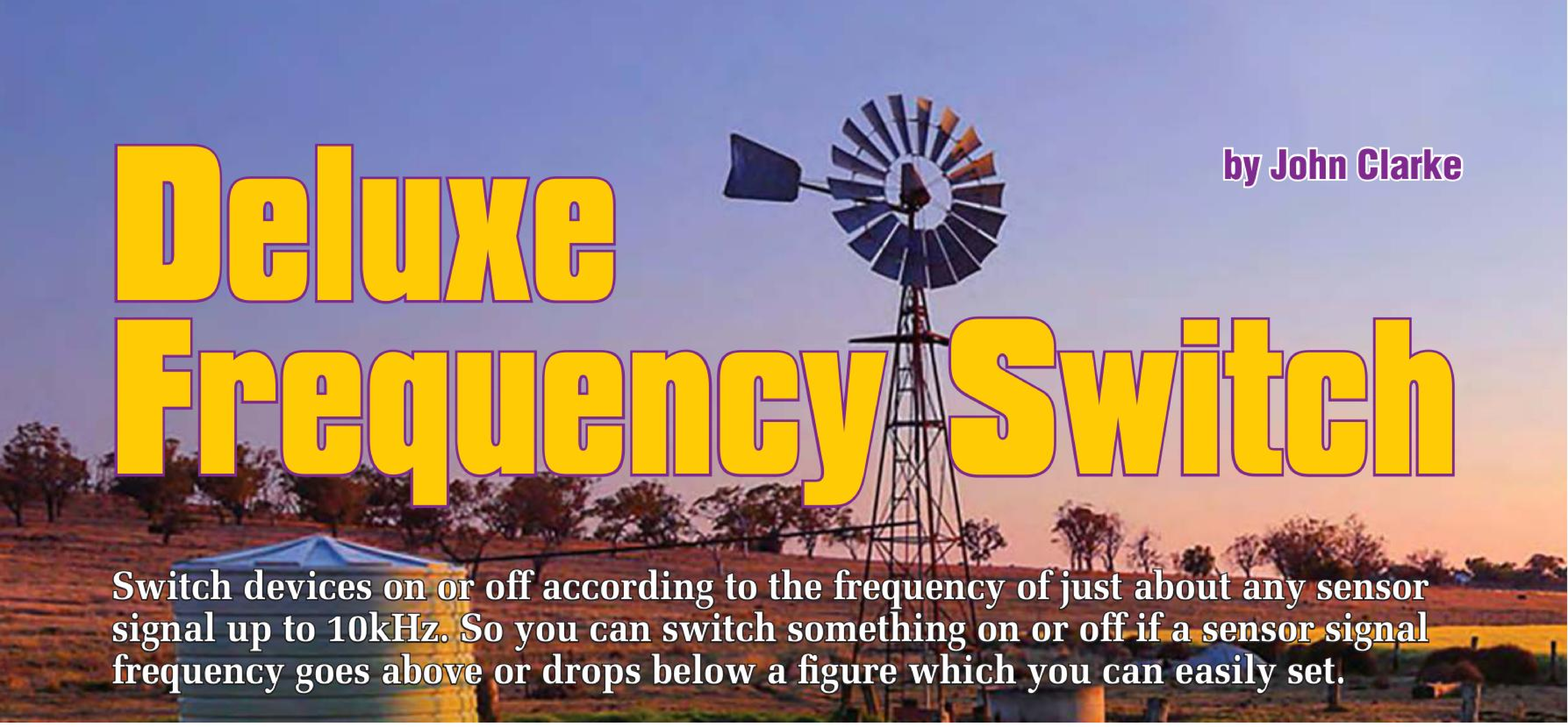
Closing date

The closing date for this offer is 31 May 2019.

January 2019 winners

Bob Kelly and
Brian Fairchild

They each won a Microchip dsPIC33CH Curiosity Development Board – valued at $35.00



Deluxe Frequency Switch

by John Clarke

Switch devices on or off according to the frequency of just about any sensor signal up to 10kHz. So you can switch something on or off if a sensor signal frequency goes above or drops below a figure which you can easily set.

There are many applications for a device of this type. Just some of the things we thought of 'off the top of our heads', include:

- Cutting power (or fuel) to a motor if it exceeds a certain speed
- Switching a fan on at low vehicle speeds to provide improved cooling
- Giving a warning to change gears when the engine RPM is approaching the tacho red line
- Switching from long to short intake runners at a particular engine RPM to optimise power delivery
- Switching off a pump if a flow meter records the water flow is outside a specific range
- Switching on an alarm if wind speed exceeds a certain threshold
- Applying a brake or feathering blades on suitably equipped turbines.

Of course, there are countless other uses – you're probably thinking of others that suit your particular application.

As long as it has, or can be fitted with a sensor, to provide a frequency which varies with speed, temperature, flow or other parameter, you can use our new *Deluxe Frequency Switch*. It can do all of this and much more.

Setting the two frequencies

You need to set up two frequencies, not one as you might have initially thought.

Why do you need two frequencies? We need to set two frequencies because if the signal from your chosen sensor varies by even a small amount at close to the switching threshold, the relay would be constantly chattering on and off – not good at all. So we set an upper frequency threshold above which the sensor signal must rise before the relay switches on. And then we set a lower frequency threshold below which the sensor signal must drop before the relay is switched off.

You can set these two frequencies close together or far apart.

Setting the frequencies is dead-easy and there are several methods for doing it. The first method is to feed in your wanted set frequency, say 500Hz, from an oscillator or other source to the sensor input and then press switch S2. Then feed in the wanted lower frequency, say 400Hz, and then press S1.

The second method is arguably even easier. You just set one frequency, say 500Hz, and then use an on-board trimpot (VR1) to set the hysteresis. This will effectively set the lower frequency (down to a minimum of 250Hz in this example) and you can tweak it at the time of installation.

If you don't have an oscillator you could use the real signal that you intend controlling the unit with, so long as you can hold it steady at the required frequency/frequencies for long enough to press the switch(es).

Alternatively, if that's too difficult, you actually can get the microcontroller to generate the wanted frequencies. This second method is more involved than the first and we will describe the procedure later in this article.

Specifications

Supply voltage	10-16V
Supply current	20mA (relay off); 60mA (relay on)
Signal frequency range	1Hz to 10kHz
Signal amplitude	>1.4V peak-to-peak
Threshold setting resolution	20Hz at 10kHz; 1Hz at 2.27kHz; 0.2Hz at 1kHz; 0.002Hz at 100Hz.
Hysteresis	0-50%
Switching delay	signal period plus 0-500ms
Signal frequency bands:	<10Hz, 10-100Hz, 100Hz-1kHz, 1-10kHz

Detection time and delay

You can also configure the unit with a switching delay which is adjusted with trimpot VR2 and can be set between zero and 500ms (ie, half a second). This ensures that if the signal frequency only momentarily crosses one of the thresholds, it will not cause the relay to switch.

The input signal frequency must remain at or beyond the threshold for the entire delay time before any relay switching will occur.

Each time the frequency crosses the threshold, the delay time starts again.

If you prefer switching to happen immediately then set the response time to zero (ie, VR2 fully anticlockwise).

LED indicators

To help in the set-up and installation procedures, we have included indicator LEDs to show when an input signal is present and its frequency range:

- LED2 lights for frequencies between 0.5Hz and 10Hz
- LED3 lights between 10Hz and 100Hz
- LED4 between 100Hz and 1kHz
- LED5 for frequencies between 1kHz and 10kHz and all four LEDs light if the frequency is above 10kHz.

Other LEDs show when the set threshold frequency is reached and whether the relay is on or off.

Relay options

The relay is a double-pole change-over (double throw) type (ie, DPDT) which can switch one or two loads, each up to 5A/48V (8A if you use the specified relay from Altronics).

You have the option to get the relay to switch on if the sensor signal rises above the threshold frequency (set by S2) and switch off if the sensor signal drops below the threshold set by S1.

The alternative is to get the relay to switch on if the sensor drops below the lower threshold frequency (set by switch S1) and switch off if the sensor signal rises above the threshold frequency (set by switch S2). The second mode is activated by installing a link at JP1 on the PCB.

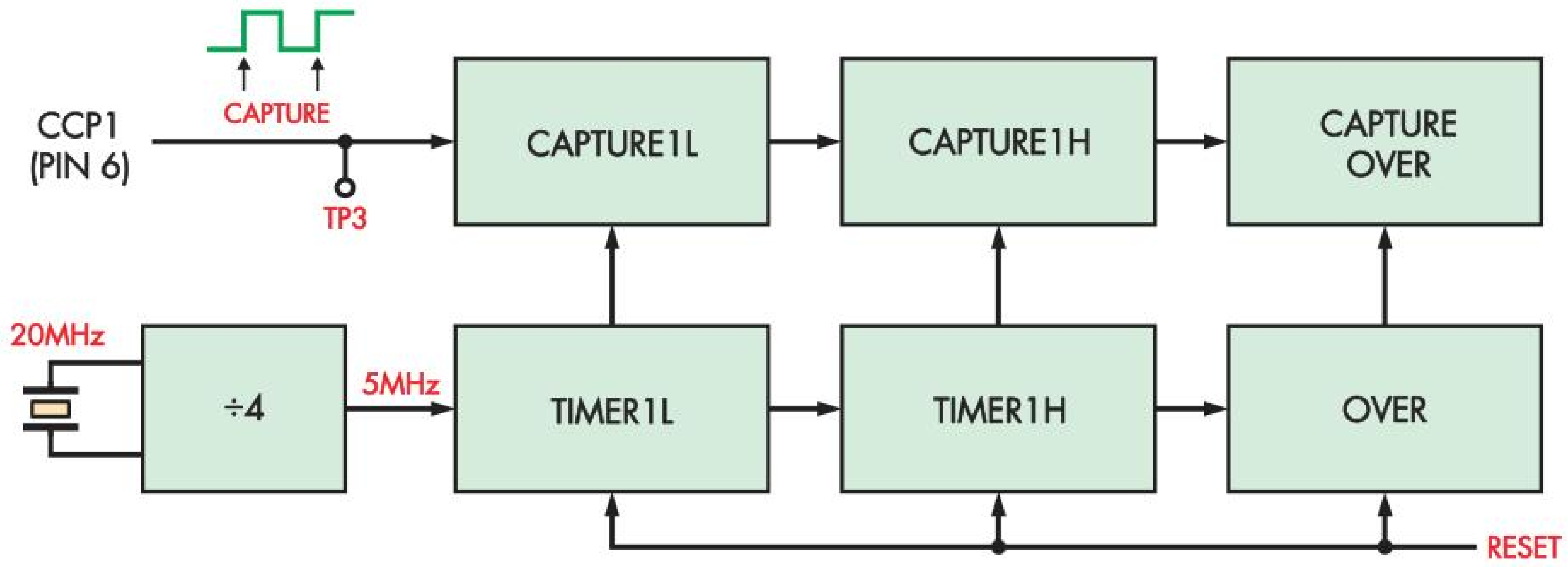


Fig.1: This block diagram describes how the microcontroller measures frequency.

Block diagram

Fig.1 (above) shows how the *Deluxe Frequency Switch* monitors the signal frequency. The PIC16F88 micro's internal clock is derived from a 20MHz crystal which is driven by an internal oscillator amplifier.

The resulting 20MHz clock signal is divided by four to produce a 5MHz signal which drives an internal 16-bit timer, TIMER1. This comprises two 8-bit cascaded timers, TIMER1H and TIMER1L.

We have implemented an 8-bit overflow counter (OVER) in the unit's firmware. That extends TIMER1 out to 24 bits, so it rolls over every 3.355 seconds [or 2^{24}÷5,000,000]. This equates to an input frequency close to 0.3Hz. Hence, the unit is designed to handle signals from 1Hz and up.

The input signal is fed to pin 6, which is also the Capture/Compare/PWM (CCP) pin. The Capture module hardware in the micro is configured so that on each positive signal transition at this pin (ie, low-to-high), the values of TIMER1H and TIMER1L are copied into the CAPTURE1H and CAPTURE1L registers and an interrupt flag is set.

This then triggers an interrupt handler routine which copies the contents of the OVER register into the CAPTURE OVER variable.

The timers and overflow counters are then reset to zero, ready to count until the next input positive-going edge.

The captured count represents the number of pulses from the 5MHz clock signal over the period between the two positive input signal edges.

So for example, a 1Hz input signal will have a one-second period between each positive edge. The count value stored will therefore be 5,000,000 (5M). At 1kHz, the period between positive edges is 1ms and the captured value will be 5000.

To calculate the frequency, all we need to do is to perform the calculation F(Hz) = 5,000,000 ÷ value. Or we can calculate the period as P(s) = value ÷ 5,000,000.

In reality, the micro just has to convert the upper and lower threshold settings to these same count units and then compare the counter values to those stored values, to determine whether either threshold has been crossed.

Features

- Energises a relay when a signal goes above a preset frequency and keeps it on until the signal drops below a second preset frequency
- Adjustable hysteresis can be used instead of setting upper and lower frequencies
- Switching frequency can be from 1Hz to 10kHz
- Adjustable switching delay
- Two sets of 5A changeover relay contacts
- Easy pushbutton set-up
- Can be set up on the bench or in situ
- Threshold can be set using a signal generator or frequency meter (eg, DMM)
- On-board signal frequency range indicators
- Power, threshold and relay-on LED indicators.

On-board frequency generation

Where the microcontroller produces an output frequency for you to measure during adjustment (as per setup method on page 23), pin 6 (CCP1) is configured differently. Rather than being in Capture Mode, with pin 6 as an input, it is used in Compare Mode and pin 6 is an output.

TIMER1 is still driven with the same 5MHz signal but the TIMER1L, TIMER1H and OVER registers are preloaded with values calculated from the frequency to be produced. Each time OVER register overflows, the pin 6 output toggles and new values are loaded into the TIMER1L, TIMER1H and OVER registers.

Because pin 6 toggles each time the counters overflow, the output frequency would be half what you might expect based on the period value for the required frequency. So we need to divide the period by two to give two separate half periods.

This means there will be an error whenever an odd period value is used, since dividing it by two will yield a remainder of one.

To solve this, and avoid the inaccuracy, two different pre-load values are used. They are used alternately to load into the TIMER1L and TIMER1H registers. So the duty cycle will not quite be 50% but the frequency produced will be accurate. The values from each of the separate period values are loaded into the TIMER1L, TIMER1H and OVER counters alternately. At the same time, the output at pin 6 is changed in level.

For those interested, the values to pre-load into TIMER1L, TIMER1H and the OVER variable are calculated as $2^{24} - (5{,}000{,}000 \div f\,(\mathrm{Hz})) \div 2$, with the alternative value being one higher in cases where 5,000,000 ÷ fHz is odd.

Circuit description

The full circuit shown in Fig.2 is based on microcontroller IC1, a PIC16F88. This monitors the input frequency, jumper state (JP1 and JP2), switch state (S1 and S2) and trimpot settings (VR1 and VR2). It also drives the frequency LEDs (LED2-LED5), threshold LED (LED6) and the relay coil (RLY1) and its associated LED (LED7) via NPN transistor Q2.

Power is fed in via CON1 and the supply is nominally 12V DC. Diode D1 provides reverse polarity protection and its cathode connects directly to the positive terminal of the relay coil, applying a nominal 11.4V to it as well as to the 5V regulator, REG1 and it powers the rest of the circuit.

A 10µF electrolytic capacitor is used to filter the supply voltage and transients are clamped using a 16V zener diode (ZD1), with the peak current limited by the series 47Ω resistor.

The supply is further filtered by another 10µF capacitor and then REG1 reduces the 11.4V supply to 5V for IC1 and input conditioning transistor Q1. The power LED (LED1) is connected across the 5V supply with a 3.3kΩ series current-limiting resistor.

The input signal is fed into CON2 and it's AC-coupled via a 10µF capacitor and 10kΩ resistor to the

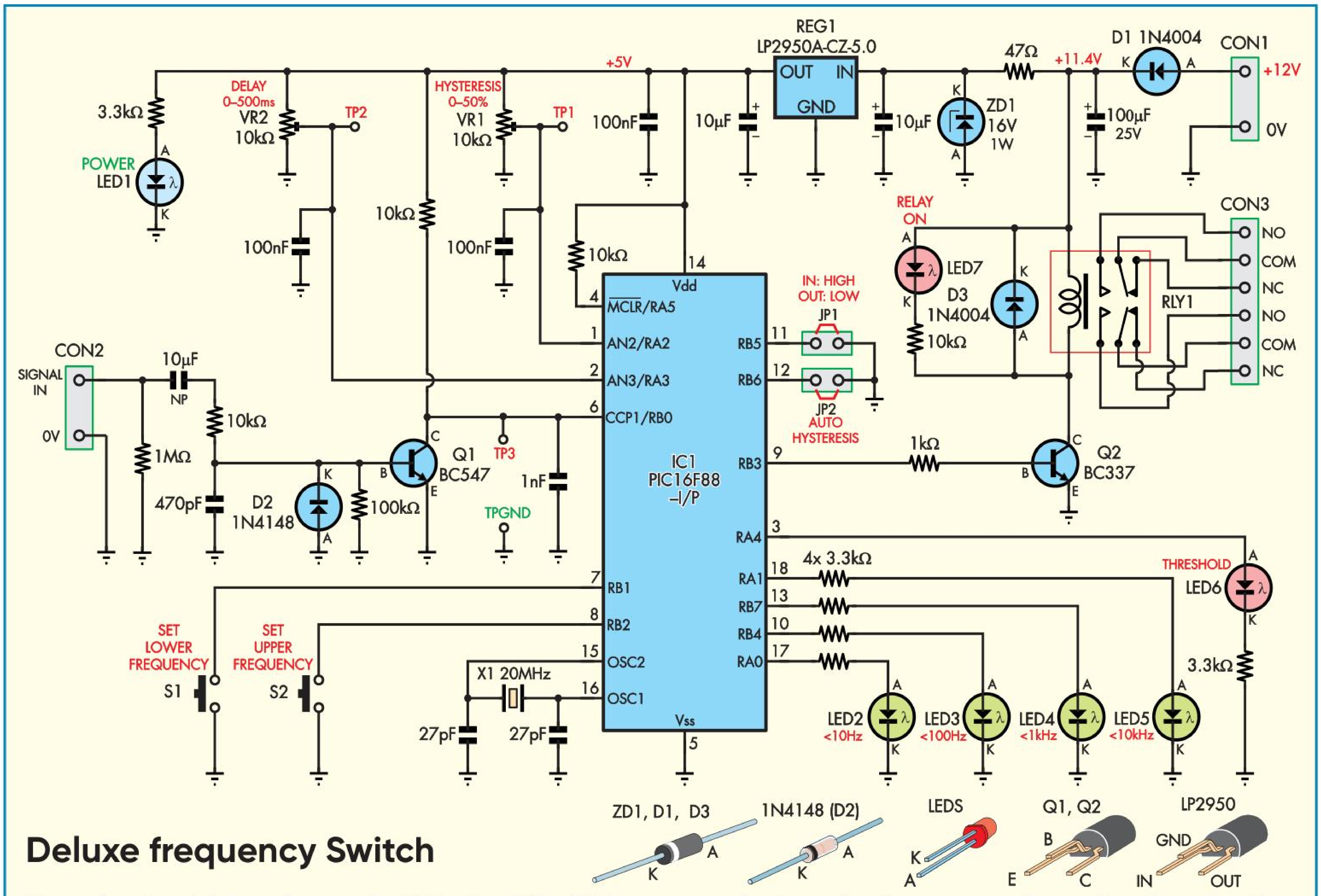

Deluxe frequency Switch
REG1
LP2950A-CZ-5.0
OUT
IN
GND
+5V
47Ω
+11.4V
D1 1N4004
K
A
CON1
+12V
0V
3.3kΩ
DELAY
0–500ms
VR2
10kΩ
TP2
HYSTERESIS
0–50%
VR1
10kΩ
TP1
100nF
10µF
10µF
ZD1
16V
1W
100µF
25V
POWER
LED1
10kΩ
100nF
100nF
10kΩ
14
Vdd
4
MCLR/RA5
1
AN2/RA2
2
AN3/RA3
6
CCP1/RB0
IN: HIGH
OUT: LOW
JP1
11
RB5
12
RB6
JP2
AUTO
HYSTERESIS
RELAY
ON
LED7
D3
1N4004
10kΩ
CON3
NO
COM
NC
NO
COM
NC
RLY1
CON2
SIGNAL
IN
0V
10µF
NP
10kΩ
1MΩ
470pF
D2
1N4148
100kΩ
Q1
BC547
TP3
1nF
TPGND
IC1
PIC16F88
–I/P
9
RB3
1kΩ
Q2
BC337
3
RA4
THRESHOLD
LED6
18
RA1
4x 3.3kΩ
13
RB7
10
RB4
17
RA0
3.3kΩ
7
RB1
8
RB2
15
OSC2
16
OSC1
X1 20MHz
27pF
27pF
Vss
5
SET
LOWER
FREQUENCY
S1
SET
UPPER
FREQUENCY
S2
LED2
<10Hz
LED3
<100Hz
LED4
<1kHz
LED5
<10kHz
ZD1, D1, D3
1N4148 (D2)
LEDS
Q1, Q2
LP2950
B
E
C
GND
IN
OUT

Fig.2: the circuit is based around a PIC16F88-I/P, which measures the incoming frequency and energises the relay if the frequency is above or below certain values and whether JP1 is present or not. It also has pre-settable response times and hysteresis to prevent 'chattering'. LEDs give you visual indication of the operation as well.

base of Q1. The 470pF capacitor filters any transients while diode D2 clamps the base voltage at –0.7V for negative excursions. Q1 inverts and amplifies the signal, suitable for the capture compare input (CCP1) at pin 6 of IC1.

Frequency measurement modes

When the micro is configured to generate frequencies for setting the upper and lower thresholds, the output signal appears at pin 6 and TP3. For this to work, there must be no input signal at CON2 and this means that Q1 is biased off and it will not load the output signal from pin 6.

20MHz crystal oscillator X1 is connected to IC1, between its CLKO and CLKI pins, to allow for accurate and wide-ranging frequency measurements. The MCLR reset input is tied to the 5V supply via a 10kΩ resistor to provide a power-on reset for the microcontroller (IC1).

Internal pull-up currents within IC1 hold the RB1 and RB2 inputs high when switches S1 and S2 are not pressed and similarly, are enabled for the RB5 and RB6 inputs which are connected to jumpers JP1 and JP2. These inputs are pulled low if a switch is pressed or jumper plug inserted, and this can be sensed by IC1.

Output pins RA0 (17), RB4 (10), RB7 (13) and RA1 (18) drive signal indicators LED2-LED5 via 3.3kΩ current-limiting resistors at around 1mA each. Similarly, output RA4 (pin 3) drives the threshold LED (LED6). The RB3 output (pin 9) switches transistor Q2 on when it goes high. This transistor in turn switches on the relay. Diode D3 quenches back-EMF from the coil as Q2 is switched off.

LED7 is also switched on when the relay is powered. It's wired across the relay coil and uses a 10kΩ series resistor due to the higher voltage (11.4V). It provides the same current to LED7 as for the other LEDs.

Trimpots VR1 and VR2 set the default hysteresis and delay time and both are connected across the 5V supply, with their wipers connected to analogue inputs AN2 (pin 1) and AN3 (pin 2) respectively. The voltages at these pins are converted to digital values using IC1's inbuilt 10-bit analogue-to-digital converter (ADC). The 100nF capacitors between each of these two pins and ground provide a low-impedance source for the ADC during conversions.

Construction

The *Deluxe Frequency Switch* is built on a double-sided PCB coded 05104181, available from the *Practical Electronics PCB Service.* It measures 102 × 58.5mm and fits in a plastic utility box measuring 129 × 68 × 43mm.

Follow the overlay diagram, Fig.3, when installing the parts. Fit the resistors first – we recommend you use a digital multimeter (DMM) to check the values before soldering them.

Diodes D1, D2, D3 and ZD1 are next, and these need to be inserted with the correct polarity, with the striped end (cathode, k) oriented as shown in the overlay diagram. Diode D2 is the 1N4148 type while D1 and D3 are 1N4004.

We recommend using an IC socket for IC1. Take care with orientation when installing the socket and when

Table 1: Hysteresis setting versus voltage at TP1.

TP1 voltage (adjusted with VR1)	**Hysteresis** when setting upper threshold
5V	50%
3.75V	43%
2.5V	33%
1.25V	20%
625mV	12%
312.5mV	6%

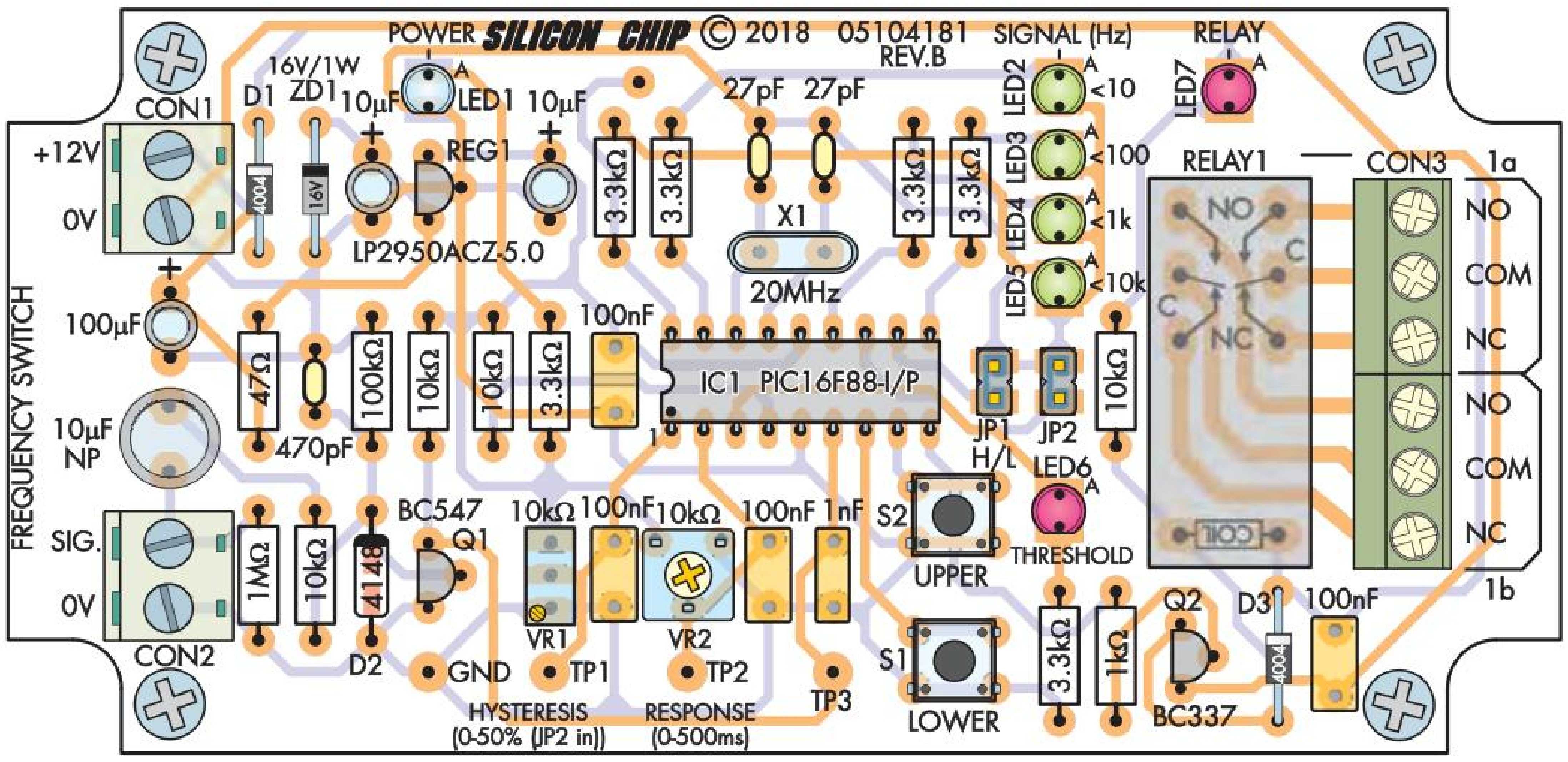
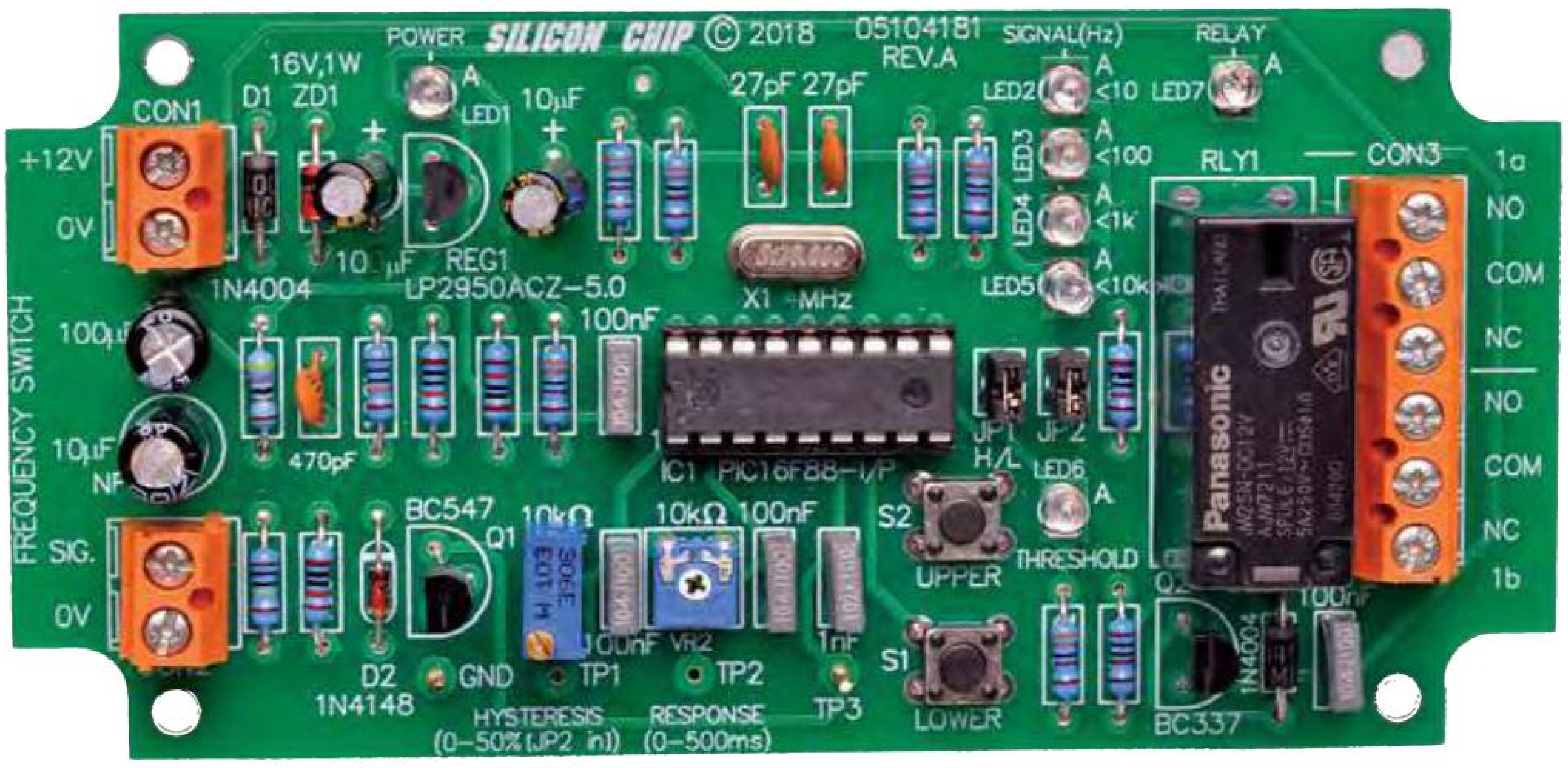


Fig.2: component layout for the *Deluxe Frequency Switch* with a matching photo below. We suggest using an IC socket for IC1 – and make sure when you place the connectors, their wiring access holes all point to the outside of the PCB.



inserting the IC. For the test points, we used two PC stakes, one for TP GND and the other for TP3. We left the remaining test points as bare pads so a multimeter probe can be inserted.

Install the two 2-way pin headers for JP1 and JP2 and then follow with the capacitors.

The electrolytic types must be fitted with the polarity shown (long lead to pad marked plus; the stripe indicates the negative side) and note that the 10µF NP capacitor is non polarised and so can be installed either way around.

Next, mount transistors Q1 and Q2 and also REG1. Take care not to mix them up because they come in identical packages.

Trimpots VR1 and VR2 are next to be fitted. They may be marked as '103' instead of 10kΩ. Orient VR1 with the adjusting screw as shown.

CON1 to CON3 can now be installed. CON1 and CON2 are 2-way types and CON3 comprises two 3-way screw connectors dovetailed together. Fit all connectors with the wire entry to the edge of the PCB.

Finally, the LEDs and relay RLY1 can be installed. We placed the LEDs close to the PCB, but they can be mounted higher or mounted off the PCB if you wish, connected with flying leads.

Although presented as a bare PCB, the unit fits in a UB3 Jiffy box. In this case, attach the PCB to the base of the box using spacers. First, mark out and drill 3mm holes for each of the corner mounting holes.

You will also need to drill holes at each end of the box for cable glands. A gland at one end is used for the power and signal wires while a gland at the other end allows the relay contacts to be wired up as required.

Set up

You have several options for setting the unit up. You can set it up before installation using an oscillator or the actual signal source (if it can be held steady enough) when you install it.

1) Oscillator method

Power the unit up with a 12V power supply wired to CON1. Connect the oscillator to CON2. Set the signal amplitude to 2V peak-to-peak or 0.7V RMS.

Set the oscillator to your desired upper threshold frequency (eg, 500Hz) and press S2. Then reduce the oscillator to set the lower threshold (eg, 400Hz) then press S1. That's all there is to it.

If you want to set a single threshold frequency (ie, the upper threshold) and use the hysteresis setting, fit a link to JP2. Then adjust trimpot VR1 for the required hysteresis (percentage) while you monitor the voltage at TP1. Then set the oscillator for the desired frequency and press S2.

Alternatively, if you want to set a single threshold frequency at the lower threshold and use the hysteresis setting for the upper threshold, fit a link to JP2. Then adjust trimpot VR1 for the required hysteresis while you monitor the voltage at TP1. Then set the oscillator for the lower threshold frequency and press S1.

Table 1 shows example relationships between the voltage at TP1 and the percentage hysteresis. For example, if you set VR1 to give 1.25V at TP1, the hysteresis will be 20% and the resulting lower threshold frequency will be 20% lower than the frequency you set with switch S2.

Note that you can also set the unit with only one threshold frequency and that will mean the relay will latch on when the signal goes above the threshold and will stay on until the power is turned off.

To set just a single threshold frequency, set the oscillator to the desired frequency and then press S2. Then disconnect the signal from CON2 and wait until the signal LEDs all are off. Then press S1 to set the lower frequency to zero.

No link is required at JP1 if you want the relay to switch on as the frequency rises above the threshold set by S2 (and turns off when the frequency drops below that set by S1).

Alternatively, install JP1 if you want the relay to switch on as the frequency falls below the threshold set by S1 (and turn off when the frequency rises above the threshold set by S2).

2) Frequency meter method

The advantage of this approach is that you don't need an oscillator but you will need a frequency meter or oscilloscope to measure the frequency at TP3.

To get into this mode, connect your frequency meter or DMM between TP3 and GND.

Switch off power, hold down both S1 and S2 and then switch on the power. The micro then produces a 100Hz signal at TP3.

To adjust this default frequency to obtain your desired upper threshold, press S1 and S2 until it reaches your target. S2 increases frequency, while S1 decreases frequency.

Short presses of the switches will alter the frequency at a slow rate. For faster changes, hold the switch down and the rate will change to a faster rate after two seconds. Continue to depress the switch for another two seconds and the frequency will change at an even faster rate.

This allows you to run through the entire frequency range in less than one minute but still be able to do finer adjustments with brief switch presses.

Having reached your target frequency, insert JP1. Then press S2 to set the upper threshold frequency. Then remove JP1. Then press S1 to reduce the frequency to the lower threshold. Then re-insert JP1 and press S1 to set the lower frequency threshold.

Note the two-step process to set each frequency. In other words,with JP1 out, use S2 and S1 to adjust the frequency to the wanted value, insert JP1 and press S2 to set the upper threshold.

Parts list – Deluxe Frequency Switch

1 double-sided PCB coded 05104181, 102 x 58.5mm
1 DPDT 12V DC coil relay (RLY1) [Jaycar SY-4052 {5A}, Altronics S 4270A {8A}]
2 2-way screw terminals with 5.08mm pin spacing(CON1,CON2)
2 3-way screw terminals with 5.08mm pin spacing (CON3)
2 2-way pin headers with shorting blocks (JP1,JP2)
1 18-pin DIL IC socket (for IC1)
1 20MHz crystal (X1)
2 SPST PCB-mount tactile pushbutton switches (S1,S2) [Jaycar SP0600, Altronics S 1120]
2 PC stakes (TP GND,TP3)

Semiconductors
1 PIC16F88-I/P microcontroller programmed with 0510418A.HEX (IC1)
1 LP2950ACZ-5.0 low dropout regulator (REG1)
1 BC547 100mA NPN transistor (Q1)
1 BC337 500mA NPN transistor (Q2)
1 16V 1W zener diode (1N4745) (ZD1)
2 1N4004 1A diodes (D1,D3)
1 1N4148 signal diode (D2)
7 3mm LEDs (LED1-LED7)

Capacitors
1 100µF 25V PC electrolytic
2 10µF 16V PC electrolytic
1 10µF non-polarised (NP) PC electrolytic
4 100nF 63V/100V MKT polyester
1 1nF 63V/100V MKT polyester
1 470pF ceramic
2 27pF NP0/C0G ceramic

Resistors and Potentiometers
(all 1%, 0.25W)
1 1MΩ 1 100kΩ 4 10kΩ
6 3.3kΩ 1 1kΩ 1 47Ω
1 10kΩ vertical multi-turn trimpot, 3296W style (VR1)
1 10kΩ mini horizontal trimpot, 3386F style (VR2)

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Then remove JP1, use S1 and S2 to obtain the lower threshold frequency, insert JP1 again and press S1.

Alternatively, if you just want to set the upper threshold frequency with S2 and have the hysteresis setting made for the lower threshold as set by trimpot (VR2), then you must have a link fitted to JP2 before you start the procedure. Similarly, you can set the lower threshold with S1 and have the upper threshold set by the hysteresis percentage value as set by VR2.

Now turning off the power takes the micro out of the mode whereby it produces an output frequency at TP3. It then reverts to normal operation, monitoring the input frequency instead.

Then fit a link to JP1 if you want the relay to switch on as the frequency falls below the threshold set by S1 (and turns off when the frequency rises above that set by S2).

No link is required at JP1 if you want the relay to switch on as the frequency rises above the threshold set by S2 (and turns off when the frequency drops below that set by S1).

If you want to configure the unit with a switching delay (as described earlier) you need to adjust trimpot VR2. You can set the delay anywhere between zero and half a second. If you don't want a delay set VR2 fully anti-clockwise.

Installation

Connect the 10-16V DC power source between the +12V and GND inputs at CON1. For automotive installations, automotive-rated wire should be used and the +12V terminal needs to connect to the switched side of the ignition.

That way, the unit only operates when the ignition is switched on and the vehicle battery won't go flat after long periods of being parked.

Using a tacho signal

Imagine you are using the engine tacho signal to switch the relay if a certain engine RPM is exceeded – for example, 6000 RPM.

If you have a 4-cylinder, 4-stroke engine, 6000 RPM = 100 revolutions per second. Since this type of engine fires two cylinders per crankshaft rotation, then the threshold should be set to 200Hz [100 x 2].

The easiest way to connect the GND terminal in a vehicle is to wire it to the chassis using a crimped eyelet secured to a convenient screw terminal.

You may need to drill a separate hole in the chassis for this connection, or utilise an existing earth connection.

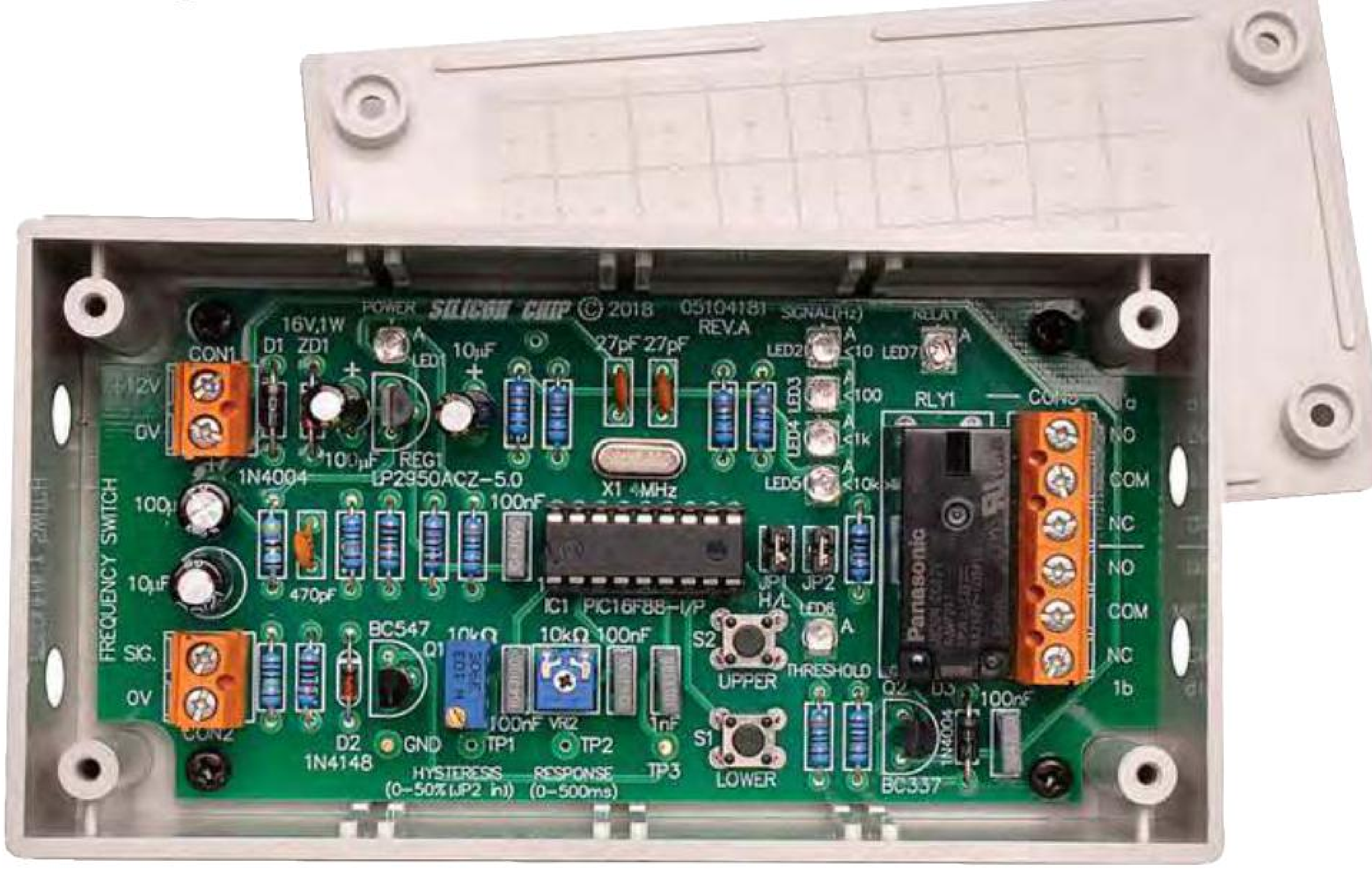
Wire CON2 to a suitable sensor. This can be the speedometer sensor, an ECU tachometer output, an injector or camshaft position sensor and so on. If you haven't already set the unit up, do so as described above above.

The relay contacts are labelled Normally Open (NO), Normally Closed (NC) and Common (COM).

To switch power to a load, wire one of its supply lines in series between either the COM and NO terminals (so that it's only powered when the relay is energised) or COM and NC terminals (so that it's switched off when the relay is energised).

Note that the relay contact current rating is 5A for the Jaycar relay and 8A for the Altronics relay (see parts list).

If a higher current is required, you can switch 12V DC to the coil of a larger relay using RLY1.



The *Deluxe Frequency Switch* PCB is designed to fit into a UB3 Jiffy box, as shown here – but it could also be 'built in' to equipment it is controlling. You may also be able to source the 10-16V DC from that equipment – as long as it isn't turned off by the frequency switch!

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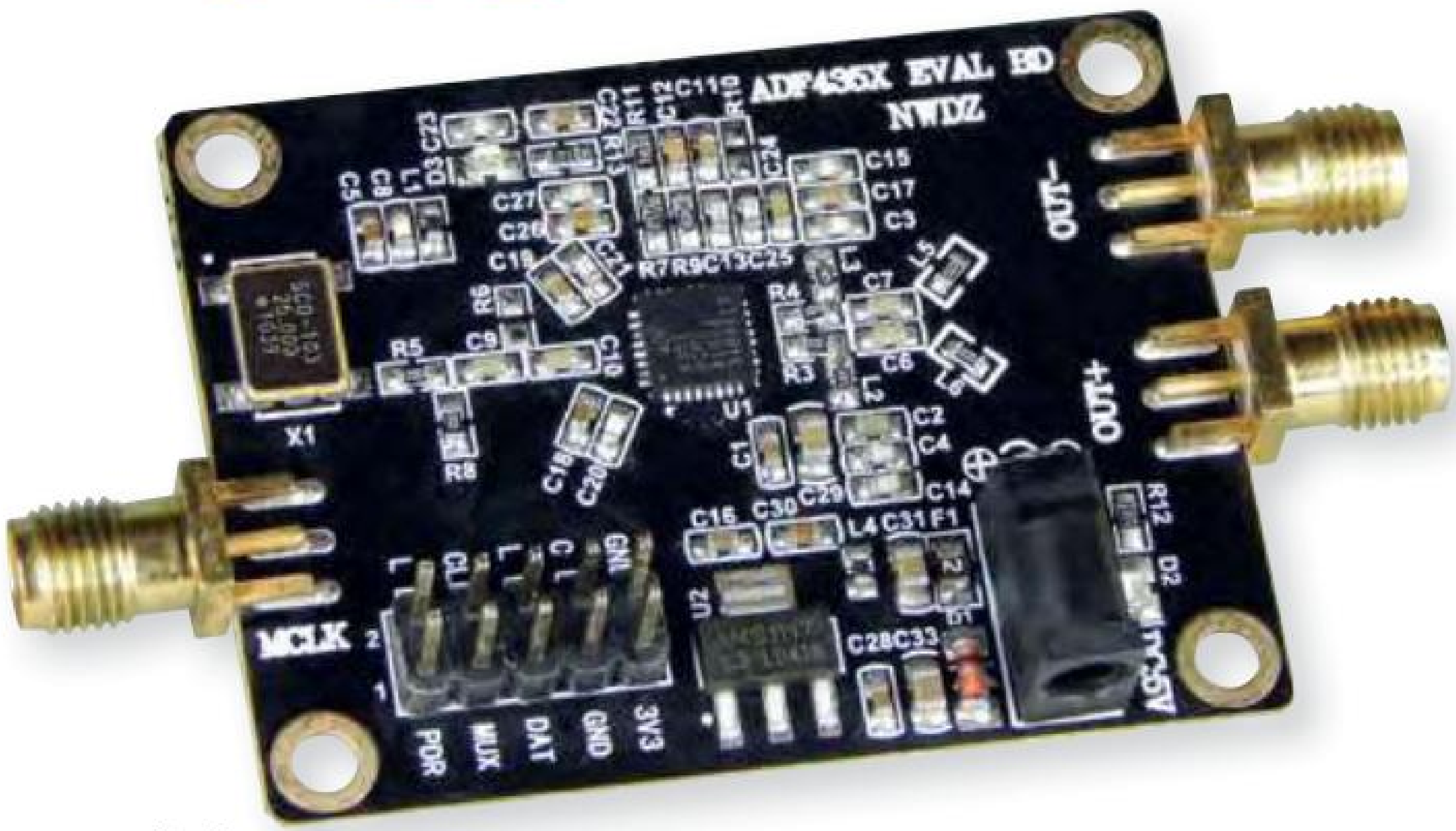
Using Cheap Asian Electronic Modules Part 16: by Jim Rowe

35MHz-4.4GHz digitally controlled oscillator



This programmable frequency module is based on the ADF4351 PLL (Phase-Locked Loop) IC and it can produce a sinewave from 35MHz to 4.4GHz, with crystal accuracy. It can even be used as a sweep generator and costs around £12.50.

That's an impressive range of frequencies that can be produced by this surprisingly compact (48 x 36.5 x 10mm) module. It is available from various Chinese websites including Banggood (**http://bit.ly/pe-may19-rf**), AliExpress and eBay, for around £12.50.

It's essentially a smaller, lower-cost version of the ADF4351 development board sold by Analog Devices. It runs from 5V and has two RF outputs, one 180° out of phase with the other, allowing it to produce either single-ended or differential signals.

It's controlled using a serial bus that's connected via a 10-pin header, which also makes connection to the 3.3V supply rail.

The ADF4351 chip at the heart of the module is an advanced phase-locked loop (PLL) device. Before we delve into how the ADF4351 works, it's a good idea to briefly cover the operation of PLLs.

Fig.1 shows the block diagram of a basic PLL. It incorporates a negative feedback loop, similar to the one used to improve the performance of audio amplifiers. However, in this case, rather than having a voltage divider providing the feedback signal, we have a frequency divider in the loop.

The PLL's output signal (F_{OUT}) is produced by the voltage-controlled oscillator (VCO) at upper right. The frequency divider divides this output frequency by a factor of N.

The resulting signal (F_{FB}) is then fed to the negative input of phase detector PD, which compares its frequency and phase with F_{REF}, the signal from a low-frequency reference oscillator, fed to its positive input.

The PD output 'error' pulses are fed to charge pump CP, which uses them to develop a fluctuating DC voltage with a polarity and amplitude proportional to the frequency/phase differences between F_{REF} and F_{FB}. This voltage is then low-pass filtered and used to control the VCO's frequency.

This feedback action causes the VCO frequency (F_{OUT}) to stabilise at very close to N times the reference frequency, F_{REF}. The PLL is then described as being 'in lock', since the feedback action keeps F_{FB} locked to F_{REF} in both frequency and phase.

So even if F_{REF} is fixed, by changing the division ratio N, we can control the frequency of F_{OUT}. Basic PLLs like this have been in use for many decades but more elaborate versions have also been developed, to overcome some of the limitations of a basic PLL.

One of these limitations is that the minimum change in F_{OUT} is equal to F_{REF}, so you need quite a low reference frequency to have fine control over the output frequency.

But it's easier to produce accurate and stable reference oscillators at higher frequencies, so one of the first enhancements to PLLs was to add a reference frequency divider between the F_{REF} input and the phase detector PD.

Also, if the output frequency needs to be up in the GHz (gigahertz) range, it's not easy to provide a programmable divider working at these frequencies.

So another early PLL improvement was to add a fixed 'prescaler' to the feedback loop, between the VCO output and the input of the main (programmable) feedback divider.

Unfortunately, this reduces the output frequency adjustment resolution. However, this can be overcome by adopting what's referred to as a 'dual modulus prescaler'.

This is essentially a prescaler with a division ratio that can be switched from one value (say P) to another (like P+1) by an external control signal.

We don't have space here to fully explain the operation of modern (and quite elaborate) PLLs, but the prior description should be enough to understand how the ADF4351 works.

Inside the ADF4351

The block diagram of the ADF4351 IC (Fig.2) is somewhat more complex than the basic PLL shown in Fig.1. The VCO part of the device is labelled 'VCO CORE' and shaded pink.

There are actually four VCOs inside the core, each used to generate a different frequency range. They are all tuned

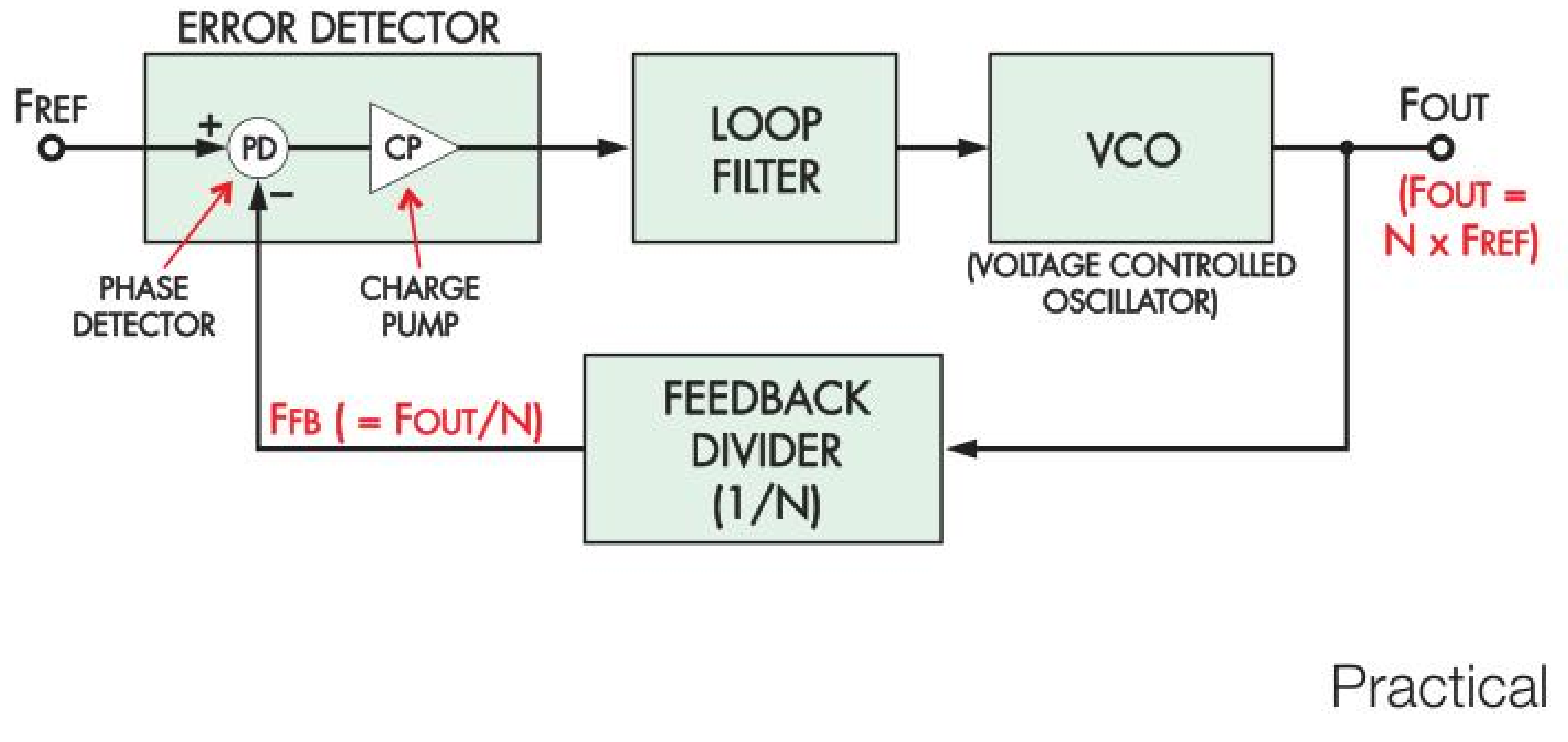


Fig.1: block diagram of a basic phase-locked loop. They're typically used to generate a stable high-frequency signal from a fixed low-frequency signal.

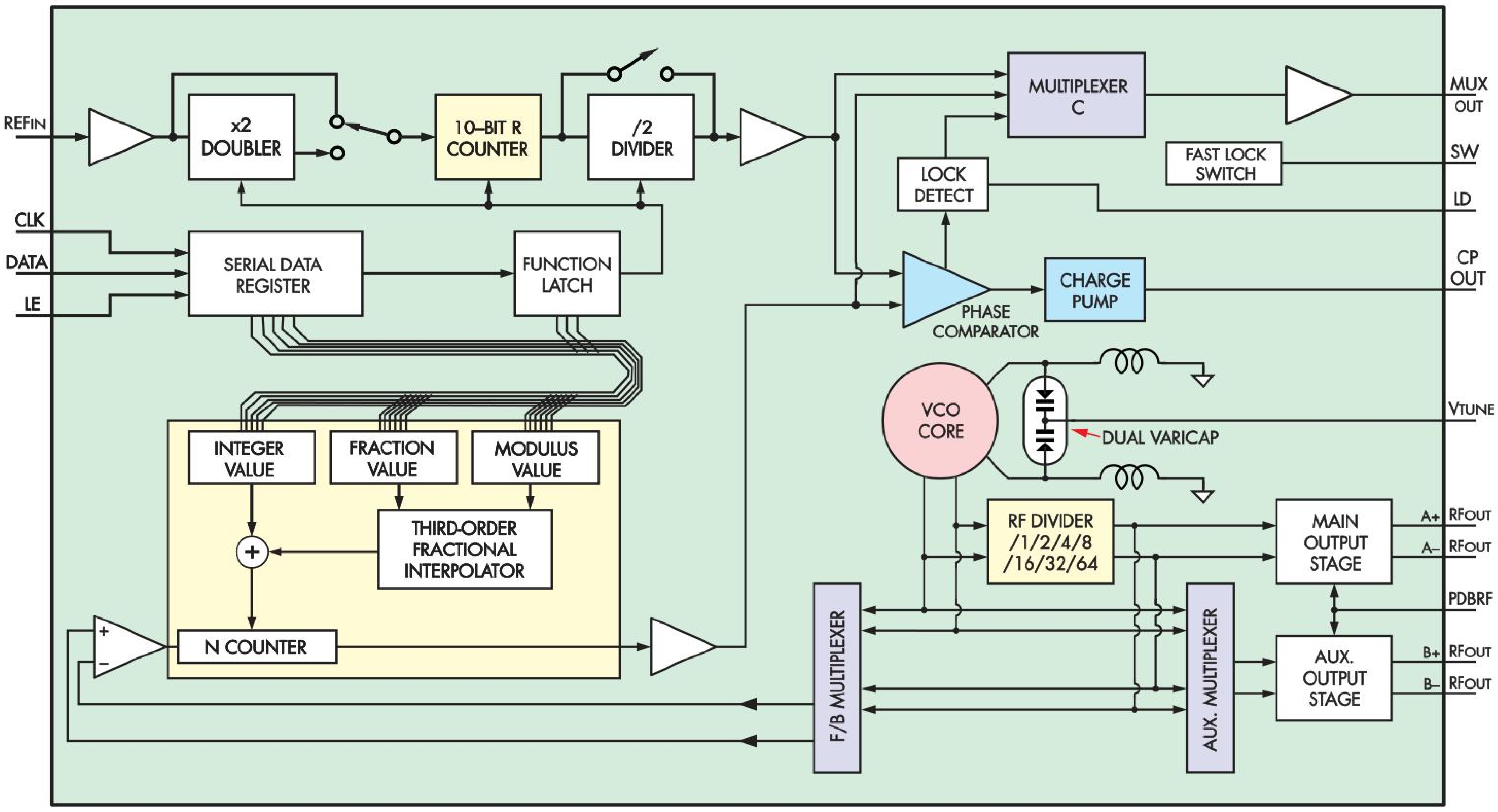

REFIN
CLK
DATA
LE
x2 DOUBLER
10-BIT R COUNTER
/2 DIVIDER
MULTIPLEXER C
MUX OUT
FAST LOCK SWITCH
SW
LOCK DETECT
LD
SERIAL DATA REGISTER
FUNCTION LATCH
PHASE COMPARATOR
CHARGE PUMP
CP OUT
VCO CORE
DUAL VARICAP
VTUNE
INTEGER VALUE
FRACTION VALUE
MODULUS VALUE
THIRD-ORDER FRACTIONAL INTERPOLATOR
N COUNTER
+
–
RF DIVIDER /1/2/4/8 /16/32/64
MAIN OUTPUT STAGE
A+ RFOUT
A– RFOUT
PDBRF
AUX. OUTPUT STAGE
B+ RFOUT
B– RFOUT
F/B MULTIPLEXER
AUX. MULTIPLEXER

Fig.2: block diagram of the ADF4351 wideband synthesiser IC. The integrated voltage-controlled oscillator has an output frequency range of 2.2 to 4.4GHz, which, when combined with the RF divider, provides the ~35MHz to 4.4GHz range. The fractional-N PLL controls the frequency from its three registers via the equation: $F_{OUT} = F_{FB} \times (INT + FRAC \div MOD)$.

by the dual varicap diode shown to its right, using a tuning voltage fed in via the V_{TUNE} pin.

Above the VCO core, you can see the phase comparator and charge pump, both blue. The charge pump output goes to the CP_{OUT} pin, so that an external low-pass filter can be used to smooth the pulsating output of the charge pump before it is fed back into the ADF4351 via the V_{TUNE} pin.

The differential outputs from the bottom of the VCO core go to three different destinations. One of these is to the yellow 'RF DIVIDER' block to its right. This programmable frequency divider can divide the VCO output frequency by 1 (ie, no division), 2, 4, 8, 16, 32 or 64.

This lets the chip generate low output frequencies while the VCO core is operating at much higher frequencies (2.2-4.4GHz).

The outputs from the RF divider are fed directly to the chip's main RF output stage, which drives the A+ out and A– out pins. The RF divider outputs also go to the inputs of two different multiplexers (digital selector switches), shown in mauve.

The auxiliary multiplexer on the right switches between the direct output lines from the VCO core and the outputs from the RF divider and so determines which is fed to the auxiliary RF output stage and then to the B+ and B– output pins. The PDBRF pin allows both RF output stages to be disabled when they are not needed, to save power.

The feedback (F/B) multiplexer at left determines which of the same two signal pairs go to the feedback divider, in the yellow box. It's also rather more complex than the simple feedback divider shown in Fig.1. That's because the ADF4351 offers the ability to implement either an integer-N or a fractional-N PLL, as required.

So the feedback divider needs three registers which hold the integer division value, the fractional division value and the modulus value, plus control circuitry labelled 'third-order fractional interpolator'. This circuitry effectively allows the feedback frequency to be divided by a rational number (fraction). The output of this divider is then fed, via a buffer, to the phase comparator.

The circuitry shown in the upper-left corner of Fig.2 takes the input from the external reference oscillator (fed into the REF_{IN} pin) and processes it before feeding it to the other phase comparator input.

As mentioned earlier, one of the refinements to earlier PLLs was to add a reference signal frequency divider, so that high-frequency reference oscillators could be used; hence the 10-bit R counter.

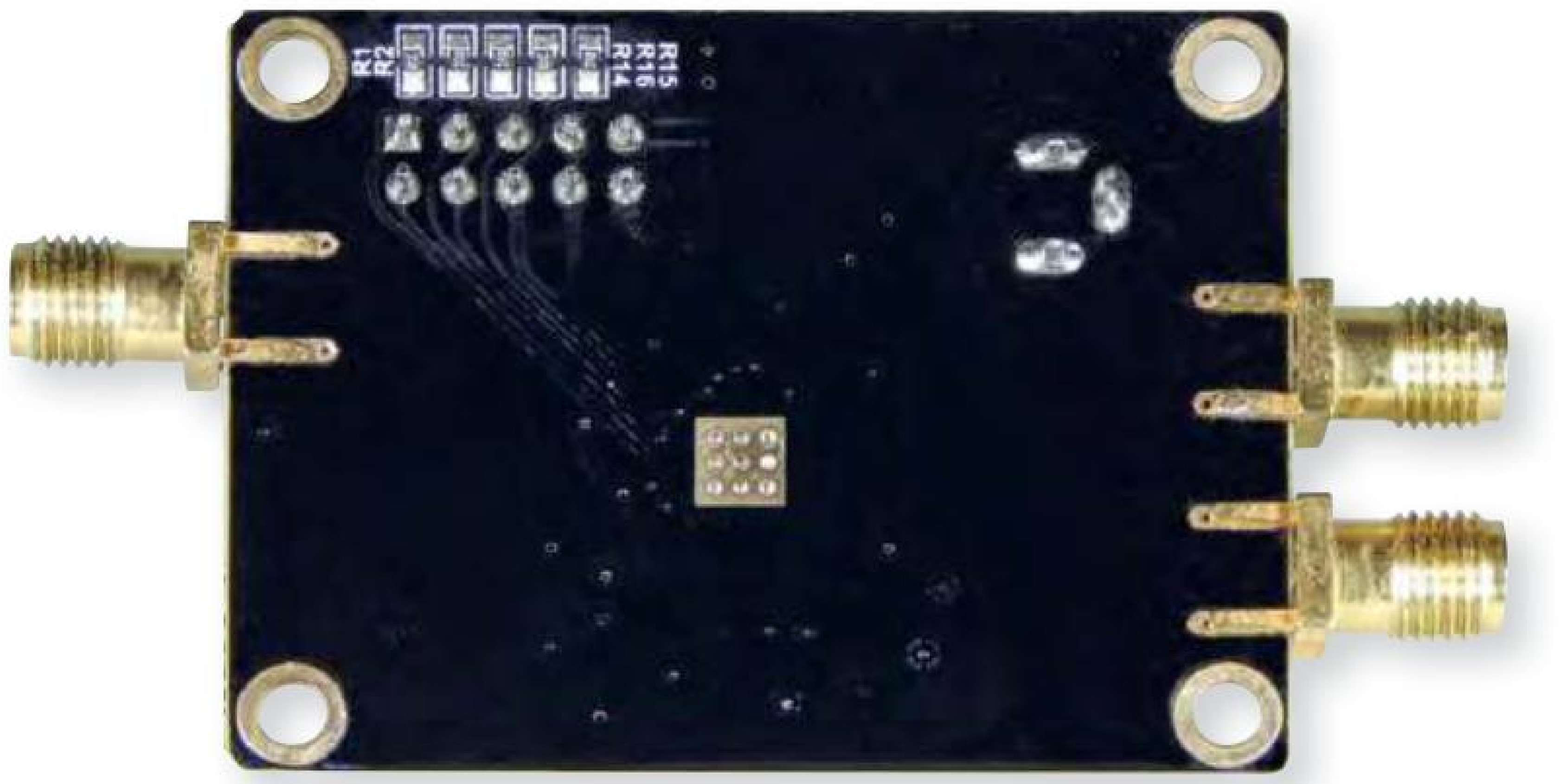
But the ADF4351 also provides a frequency doubler and an additional divide-by-two stage for the reference input, both of which can be switched in or out under software control. This gives the chip a great deal of flexibility.

The whole chip is controlled by means of a simple 3-wire serial peripheral interface (SPI), shown at centre left of Fig.2.

Serial data from the PC or microcontroller is fed in via the DATA pin, clocked into the serial data register and function latch via clock pulses fed to the CLK pin, and then latched into the various control registers by feeding in a pulse via the LE (latch enable) pin.

All functions of the ADF4351 chip are configured using six 32-bit control words, sent over this serial bus.

Multiplexer C and the other blocks at the upper right of Fig.2 allow external monitoring of the ADF4351's status.


R1
R2
R15
R16
R14

The bottom view of this module is shown at approximately twice actual size. The bottom of the board is populated by five 10kΩ pull-down resistors for the breakout pin connections.

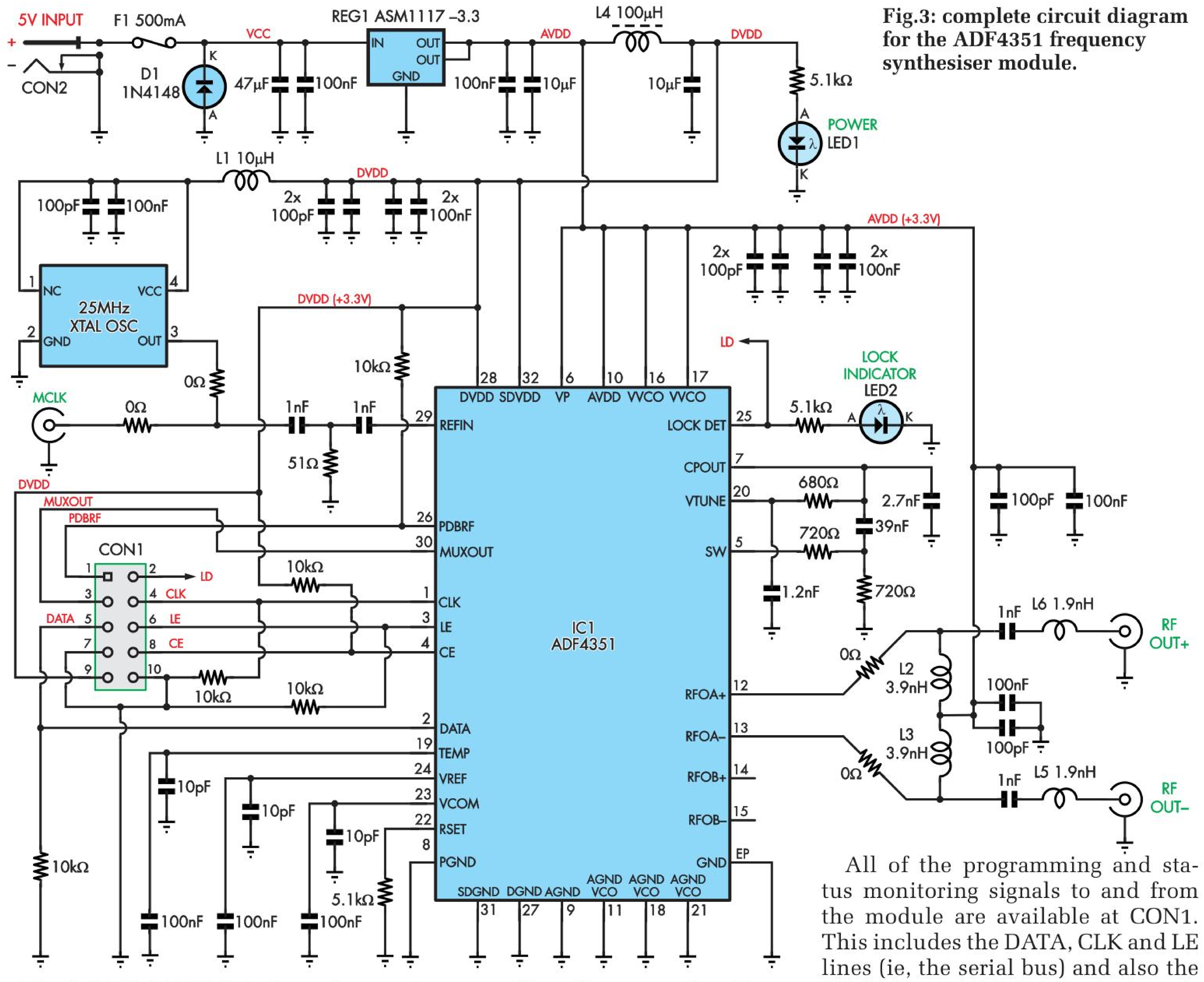

5V INPUT
F1 500mA
CON2
D1 1N4148
VCC
47µF
100nF
REG1 ASM1117 –3.3
IN
OUT
OUT
GND
AVDD
100nF
10µF
L4 100µH
10µF
DVDD
5.1kΩ
POWER LED1
L1 10µH
DVDD
100pF
100nF
2x 100pF
2x 100nF
AVDD (+3.3V)
2x 100pF
2x 100nF
NC
VCC
25MHz XTAL OSC
GND
OUT
DVDD (+3.3V)
10kΩ
0Ω
MCLK
0Ω
1nF
1nF
51Ω
LD
LOCK INDICATOR LED2
5.1kΩ
DVDD SDVDD VP AVDD VVCO VVCO
REFIN
LOCK DET
CPOUT
VTUNE
SW
680Ω
2.7nF
39nF
720Ω
720Ω
1.2nF
100pF
100nF
DVDD
MUXOUT
PDBRF
CON1
LD
CLK
LE
CE
DATA
10kΩ
10kΩ
10kΩ
PDBRF
MUXOUT
CLK
LE
CE
IC1 ADF4351
DATA
TEMP
VREF
VCOM
RSET
PGND
RFOA+
RFOA–
RFOB+
RFOB–
GND
EP
0Ω
L2 3.9nH
L3 3.9nH
0Ω
1nF
L6 1.9nH
RF OUT+
100nF
100pF
1nF
L5 1.9nH
RF OUT–
10pF
10pF
10pF
10kΩ
5.1kΩ
100nF
100nF
100nF
SDGND DGND AGND AGND VCO AGND VCO AGND VCO
31 27 9 11 18 21

Fig.3: complete circuit diagram for the ADF4351 frequency synthesiser module.

The 'LOCK DETECT' block monitors the phase comparator and provides a high logic output on the LD pin when the PLL is locked. Multiplexer C allows either of the two phase comparator inputs or this lock status to be fed to the MUX_{OUT} pin.

The fast lock switch provides a signal which can be fed into the external low-pass filter (between the CP_{OUT} and V_{TUNE} pins) when in 'fast lock' mode. So that covers the operation of the IC itself.

The synthesiser module

The full circuit of the module is shown in Fig.3, and most of the real work is done by IC1.

All of the programming and status monitoring signals to and from the module are available at CON1. This includes the DATA, CLK and LE lines (ie, the serial bus) and also the CE (chip enable), LD (lock detect), MUX_{OUT} and PDBRF (power down RF buffer) lines.

The reference signal is provided by a 25MHz crystal oscillator (XO), shown at upper left, with its output fed to the REF_{IN} pin of IC1 via a loading/coupling circuit comprising two 1nF capacitors and a 51Ω resistor.

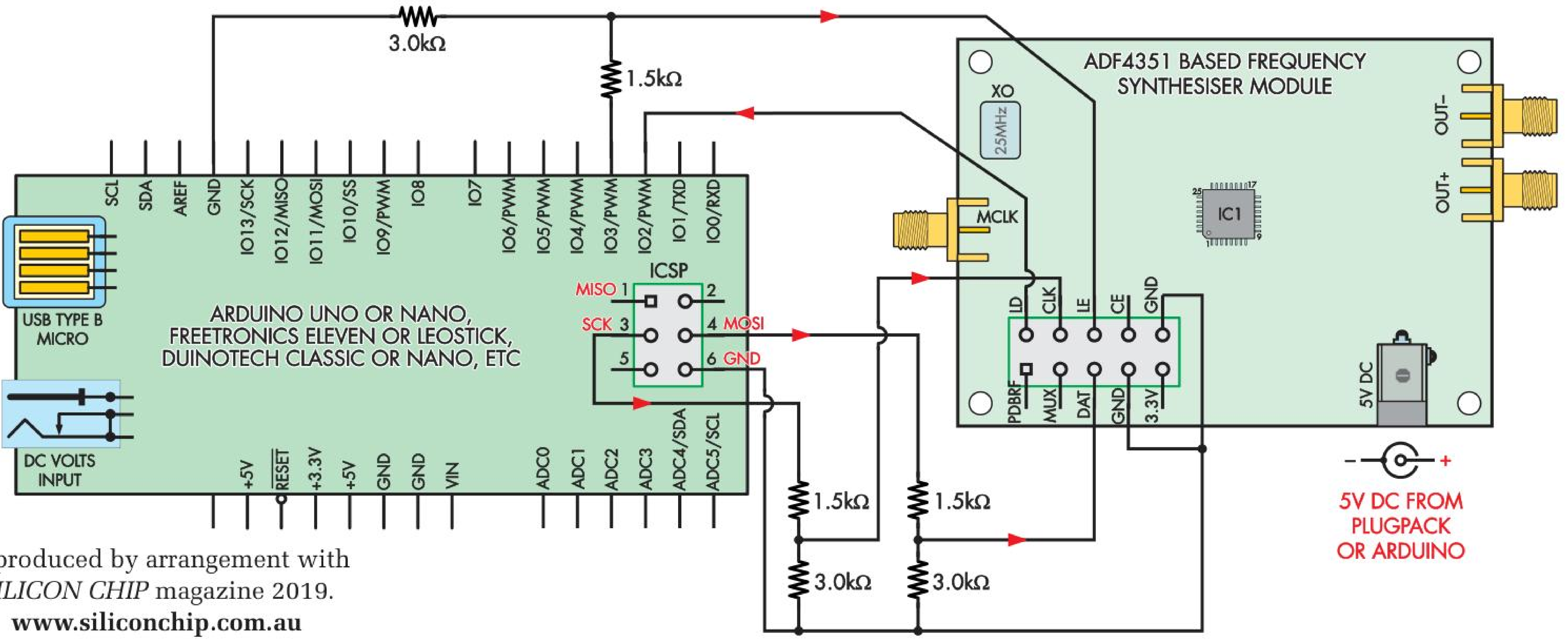

3.0kΩ
1.5kΩ
SCL
SDA
AREF
GND
IO13/SCK
IO12/MISO
IO11/MOSI
IO10/SS
IO9/PWM
IO8
IO7
IO6/PWM
IO5/PWM
IO4/PWM
IO3/PWM
IO2/PWM
IO1/TXD
IO0/RXD
USB TYPE B MICRO
ARDUINO UNO OR NANO, FREETRONICS ELEVEN OR LEOSTICK, DUINOTECH CLASSIC OR NANO, ETC
ICSP
MISO
SCK
MOSI
GND
DC VOLTS INPUT
+5V
RESET
+3.3V
+5V
GND
GND
VIN
ADC0
ADC1
ADC2
ADC3
ADC4/SDA
ADC5/SCL
ADF4351 BASED FREQUENCY SYNTHESISER MODULE
XO
25MHz
MCLK
IC1
OUT–
OUT+
LD
CLK
LE
CE
GND
PDBRF
MUX
DAT
GND
3.3V
5V DC
5V DC FROM PLUGPACK OR ARDUINO
1.5kΩ
1.5kΩ
3.0kΩ
3.0kΩ

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Fig.4: when connecting the ADF4351 synthesiser module to an Arduino-based device, a few extra resistors are needed. These resistors form a voltage divider, as the module can only handle 3.3V signals, while the Arduino's outputs have a swing of 5V. Note the changes needed if using a V2 module at the end of this article.

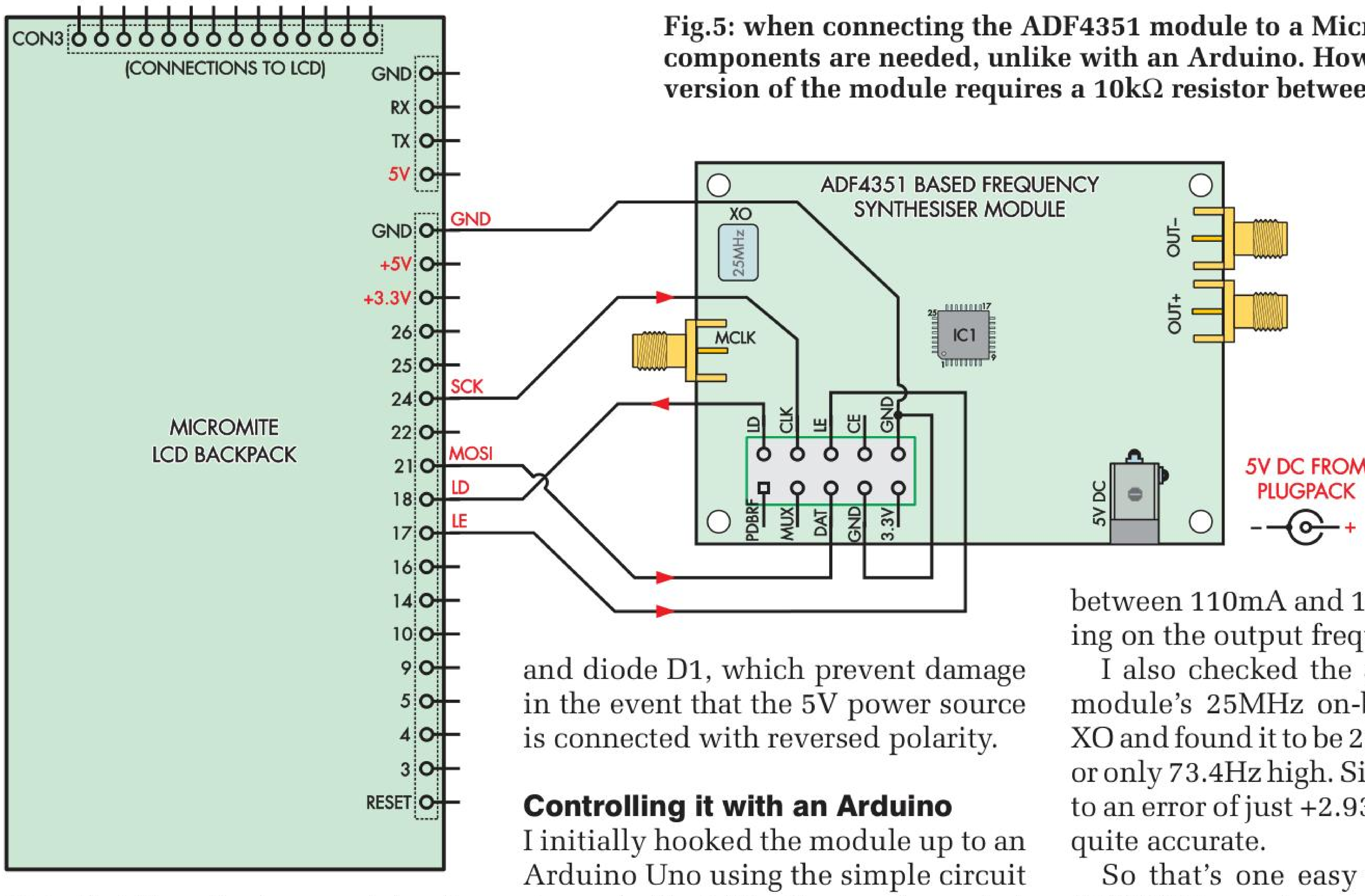


Fig.5: when connecting the ADF4351 module to a Micromite no extra components are needed, unlike with an Arduino. However, the newer version of the module requires a 10kΩ resistor between CE and +3.3V.

Note that there is also provision for feeding in a different reference signal, via the SMA socket labelled MCLK. In order to do this, you'd need to remove the 0Ω resistor connected to pin 3 of the onboard XO. If you are using the onboard XO, the MCLK socket can be used to monitor its output via a scope or frequency counter.

The resistors and capacitors connecting the CP_{OUT} and SW pins of IC1 (pins 7 and 5) to the V_{TUNE} pin (pin 20) form the low-pass loop feedback filter.

The RF output signals from the RFOA+ and RFOA– pins (12 and 13) are taken to the RF_{OUT+} and RF_{OUT-} SMA sockets via matching/filtering circuits using L2, L3, L5 and L6, plus two 1nF capacitors.

Notice that the output pins are fed with the +3.3V supply voltage via L2 and L3. In this module, the auxiliary RF outputs RF_{OB+} and RF_{OB-} (pins 14 and 15) are not wired up.

The whole module operates from a 3.3V supply, derived from the 5V input at CON2 via REG1, an ASM1117 low-dropout regulator. This AV_{DD} rail powers all of the analogue/RF circuitry directly.

The digital supply rail, DV_{DD}, is derived from AV_{DD} using LC filters comprising inductors L4 and L1, plus a number of bypass capacitors.

There are two indicator LEDs. LED1 is connected between the DV_{DD} line and ground and indicates when the module has power, while LED2 is connected to the LD (lock detect) pin of IC1 (pin 25) and indicates when the PLL is in lock.

All the other components are for bypassing and stability, apart from fuse F1 and diode D1, which prevent damage in the event that the 5V power source is connected with reversed polarity.

Controlling it with an Arduino

I initially hooked the module up to an Arduino Uno using the simple circuit shown in Fig.4. The three main control lines MOSI, SCK and LE are not taken directly to the DAT, CLK and LE pins of the module, but instead via 1.5kΩ/3kΩ voltage dividers.

This is because the inputs of the ADF4351 can only cope with 3.3V signals, whereas the Arduino outputs have a 5V output swing.

The LD signal fed back from the module to the Arduino's D2 pin does not need a divider because it's going the other way and the Arduino inputs function well with a signal having a swing of 3.3V.

Note also that Fig.4 indicates that the 5V supply for the module can come from either a plugpack or from the 5V output of the Arduino. I adapted an Arduino sketch I found on the internet, written by French radio amateur Alain Fort F1CJN (**http://bit.ly/pe-may19-af**).

Mr Fort's sketch was written for an Arduino with an LCD button shield, but I decided to adapt it so that it would work with the simple configuration shown in Fig.4, relying on the Arduino IDE's Serial Monitor to send commands to the ADF4351 and to indicate the PLL's output frequency and whether it was locked or not.

I also connected one of the PLL module's RF outputs to my frequency counter, via a prescaler, so I could monitor it.

The results were quite impressive. I could type in any frequency between 35MHz and 4.4GHz, with a resolution of 0.01MHz (10kHz) and the module's output would lock to that frequency in the blink of an eye.

I also monitored the current drawn by the module and found that it varied between 110mA and 145mA, depending on the output frequency.

I also checked the accuracy of the module's 25MHz on-board reference XO and found it to be 25.0000734MHz, or only 73.4Hz high. Since this equates to an error of just +2.936ppm, it seems quite accurate.

So that's one easy way to get the ADF4351 module going with an Arduino. The sketch (**ADF4351_and_Arduino_SC_version.ino**) is available for download from the *Practical Electronics* website.

Driving it from a Micromite

I also hooked the module up to a *Micromite LCD BackPack* combination and wrote some code so that it could be controlled via the LCD touchscreen.

The circuit is shown in Fig.5 and it's about as simple as you can get. In this case, no resistive dividers are needed on the SCK, MOSI and LE lines because the Micromite's logic pins have a swing of 3.3V.

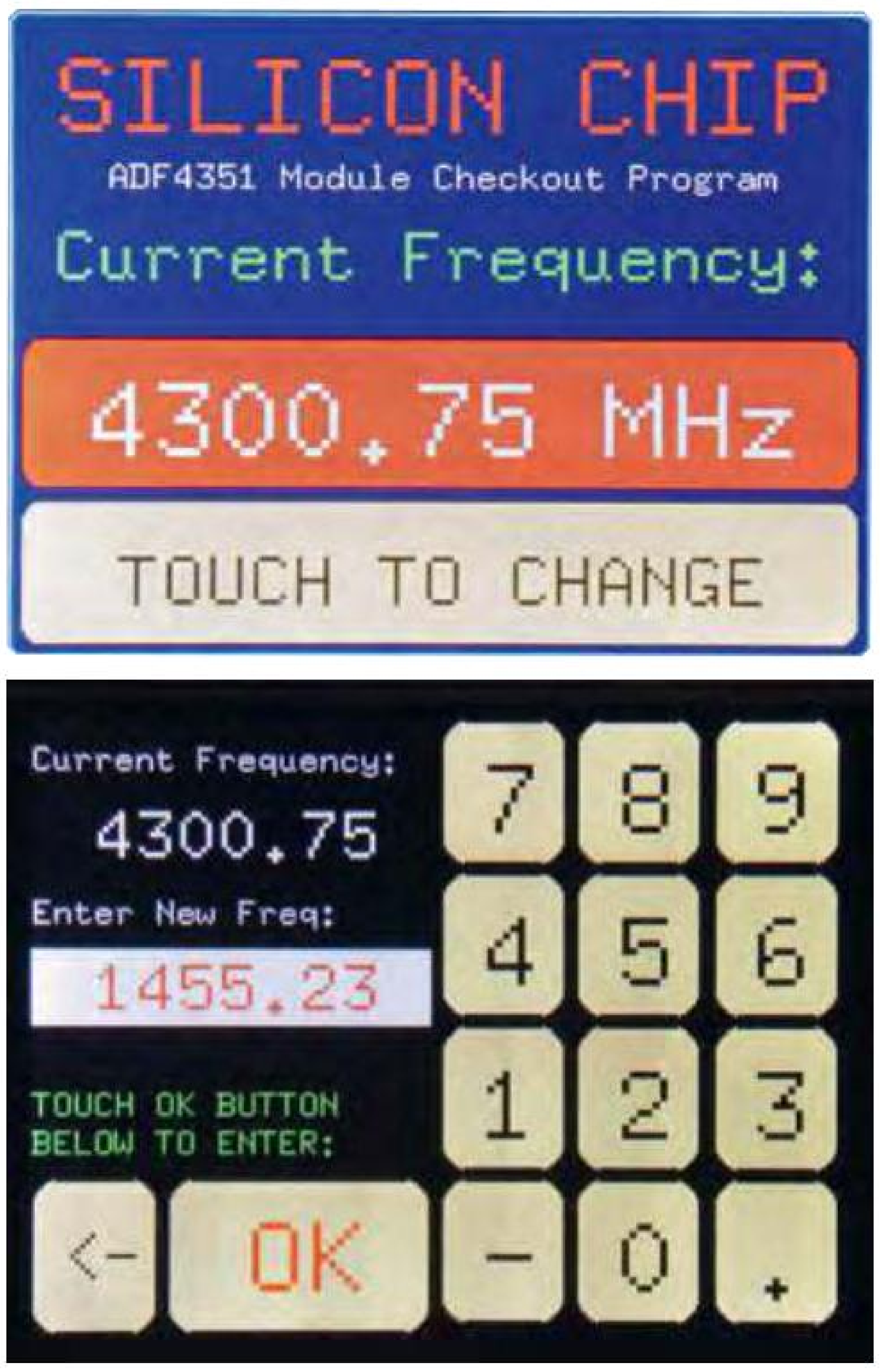
I used a 'software' SPI port rather than the hardware one used by the Micromite to communicate with the LCD and touchscreen, to prevent possible interaction.

The embedded C code (CFUNCTION) needed to provide this added port is included in the MMBasic program I wrote for this approach. Software SPI port performance is limited, but that isn't a problem as the amount of data to transfer is small.

A USB charger was used to supply 5V to the ADF4351 module because its current drain is a little too high for the *BackPack* to provide.

The software uses just two screens, as shown below. The initial screen (at left) displays the current frequency and gives you the option of touching the button at the bottom if you want to change it.

You will then get the second screen, which allows you to key in a new frequency, which is displayed below the current frequency.



The sample program running on a *Micromite LCD BackPack*. These are the only two screens the software uses: one to enter a specific frequency for the module to output and another to display the current frequency.

When you're happy with the new figure, simply touch the OK button and the module jumps to the new frequency. The program returns to the main screen, displaying the new frequency.

So for those who would like to team up the module with a Micromite, this program (**Simple ADF4351 driver program.bas**) should get you off to a good start. Like the Arduino sketch, it's available from the *Practical Electronics* website.

Performance

I checked the module's RF output performance at quite a few different frequencies, using my Signal Hound USB-SA44B spectrum analyser, which was controlled by Signal Hound's 'Spike' software.

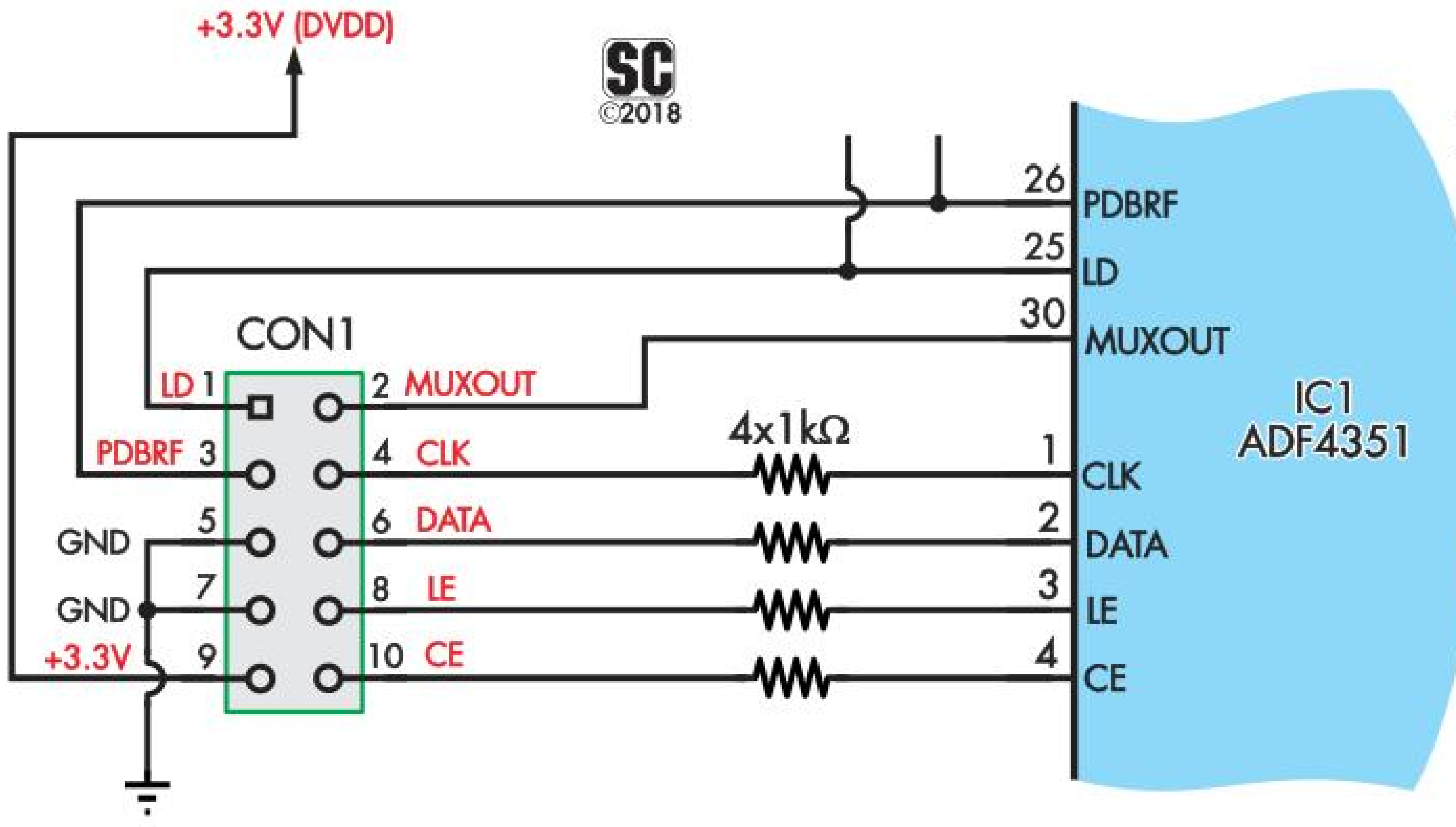


Fig.6: CON1 has been changed completely on the newer version of the ADF4351 module. Every signal, except for CLK, is connected to a different pin location.

The results were quite impressive, as you can see from the two spectrum plots. One plot shows the output at 275MHz, with the only significant spurs visible being at ±50MHz with an amplitude of –57dBm.

The other plot shows the output at 4.200GHz, with two spurs again visible but this time both on the low side: one at 4.150GHz (far left) with an amplitude of about –53dBm and the other at 4.175GHz with an amplitude of –61dBm. In both cases, the amplitude of the main output carrier is very close to 0dBm. This turned out to be the case over most of the range, in fact.

The only region where the carrier level did drop (to around –20dBm) was in the vicinity of 2.45GHz – perhaps by design, to minimise interference with Wi-Fi and Bluetooth systems.

Overall, the ADF4351 frequency synthesiser module is very impressive, especially when you consider its frequency range and price.

It could even be used to make your own VHF/UHF signal and sweep generator, teamed up with a Micromite and a 4GHz digital attenuator module that we will describe in next month's issue.

A new version of the ADF4351 synthesiser module

Just recently we received a second ADF4351 Synthesiser module and discovered that it was a 'V2' module which had been changed in a number of ways compared with the original version ('V1').

These changes will be critical to successfully connect the module to a micro, so here are the main details listed below:

1. Many of the connections to the 10-way pin header (CON1) have changed, as shown in Fig.6.
2. There is now no on-board pull-up resistor connecting IC1's CE pin to the +3.3V (DV_{DD}) line, nor are there pull-down resistors connected between the CLK, DATA and LE pins and ground.

To ensure normal operation of the module with either an Arduino or a Micromite, an external 10kΩ resistor must be connected between the CE and +3.3V pins of CON1.

To ensure maximum stability, it's a good idea to also connect an external 10kΩ resistors between the LE pin and the ground.

Once the above changes are made, version 2 of the module performs just as well as the earlier version.

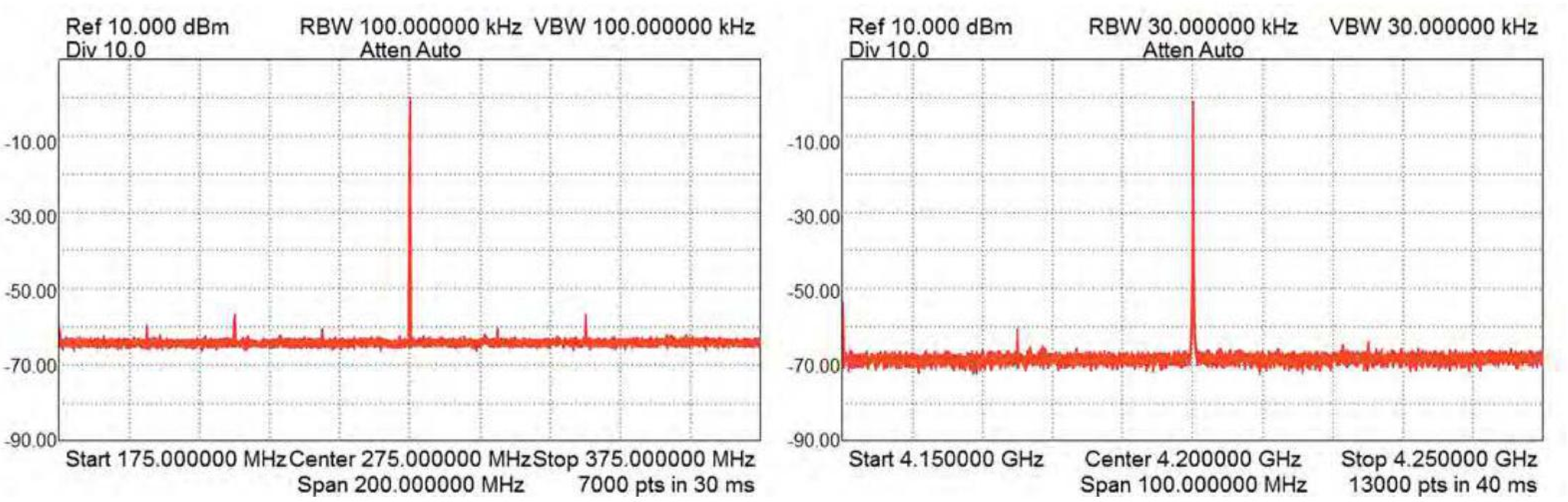
Useful links

The module, available from AliExpress: **http://bit.ly/pe-may19-ali**

The module from eBay – at the time of going to press, item 143041004110

Analog Devices ADF4351 data sheet: **http://bit.ly/pe-may19-ad**

Fundamentals of PLLs from Analog Devices: **http://bit.ly/pe-may19-pll**



Spectrum analysis of the ADF4351 module's output performance at 275MHz (left) and 4.2GHz (right). The RF output performance over the full range was good, with only a few visible spurs outside the programmed frequency. These normally correspond to beat frequencies or integer-multiples of the reference and oscillator frequency.

USB PORT PROTECTOR

LED1
1k
REF1
102
473
47k
470
CON1
LED2
PTC1
Q2
1002
10k

Don't let one small 'oops' fry your PC!

by Nicholas Vinen

Using your PC or laptop to power a 5V project that you're working on is very convenient – but it's *so* easy to make a small slip while plugging something into a breadboard... and... oops! That's exactly what happened to one of our staff members. For a while after the incident, it looked like his laptop was toast. But fortunately he was able to safely reset it and it came back to life. But he was *so* lucky! Next time he won't trust to luck – he'll definitely be using this simple, economic device.

We won't name the hapless person who thought he'd cooked his laptop. To avoid embarrassment, we'll simply refer to him as 'A.P.' (aka, 'Accident Prone').

This is one of those projects we know will be useful because A.P. kept asking 'is it finished yet' – he obviously needed it!

That incident spooked him and why wouldn't it? He could have lost a lot of work and spent quite a bit of money and time on buying a new computer and then setting it up, which could have taken several days.

We do a lot of development work, increasingly with Arduinos and similar microcontroller modules. We also do quite a bit of bread-boarding, often in combination with the Arduinos.

When you're doing this kind of work and you have external power supplies or voltage sources connected to your circuit, that's just asking for trouble.

You may not realise it, but when an Arduino board (or similar) is plugged into your computer's USB port, you're just a slip away from potential disaster.

For example, say you're running the Arduino from a 12V plugpack, because it's driving some 12V relays or a motor or whatever. So there's a source of 12V right near a bunch of other connections on the Arduino board, just looking for an excuse to find its way onto the USB 5V rail and into your computer. One slip, and oops! It could blow up the Arduino, your shield(s), and even your computer.

Not only will this *USB Port Protector* vastly improve the chances of your computer surviving such an event, it may also prevent damage to the Arduino board and whatever shields or other circuitry are plugged into the USB port. We can't promise it will be 100% safe, but it's certainly a lot safer than if you aren't using any protection...

You might expect USB ports to have some kind of built-in protection against external voltages being fed in. After all, all kinds of devices can be plugged into these ports, including external hard disks and amplifiers and other gear which may have its own, separate power supply.

In fact, many USB ports do have some kind of protection, for example series PTC thermistors ('polyswitches') to limit fault currents, transient voltage suppressors and so on. But this protection varies between computers and is often absent in laptops and notebook computers.

Let's face it, there's a lot less space inside portable computers – and manufacturers also want to keep the computer as light as possible and save money where they can. That means leaving out anything that isn't absolutely necessary.

Regardless of what sort of protection your USB port may have, this *USB Port Protector* is small and cheap, so why not add in an extra layer of defence? If you ever manage to activate its protection, it will have paid for its cost many times over!

Circuit description

The circuit of the *USB Port Protector* is shown in Fig.1. USB plug CON1, which plugs into your computer, is shown on the left side, while the USB socket, CON2, goes to the connected device (eg, Arduino) and is on the right. Just to be clear – the potential danger of overload from excessive voltages or currents comes via CON2.

The ground connection and the two differential data lines, D+ and D–, are wired straight through between plug and socket (ie, CON1 and CON2) while the 5V power flows through fuse F1 and positive temperature coefficient thermistor PTC1.

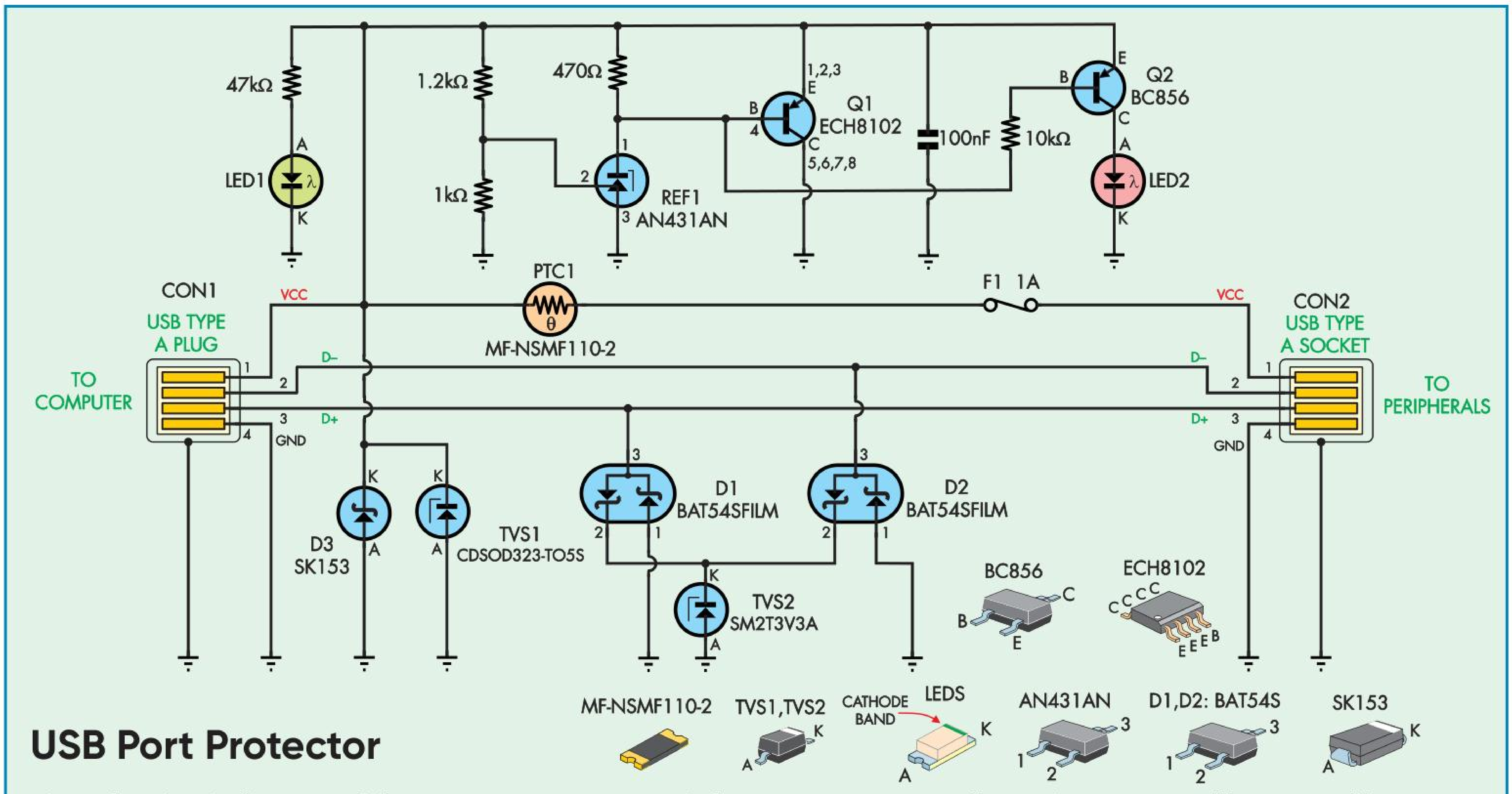


Fig.1: the circuit diagram of the *USB Port Protector*. Diode D3, zener TVS1 and transistor Q1 are all connected between V_{CC} and GND and shunt current when an excessive voltage is applied, while polyswitch PTC1 and fuse F1 prevent large currents from flowing if the fault is serious. Diodes D1, D2 and zener TVS2 protect the D+ and D– data lines.

We've used both a fuse and PTC because the fuse reacts faster to very high currents, protecting the rest of the circuitry on the board if there's a serious fault, but the PTC does not need to be replaced if it 'trips' and helps the circuit to handle moderate overloads without damage.

PTC1 normally has a low resistance – around 100mΩ below 1A – but if the current through it increases, its resistance rises, limiting it at around 2A (given enough time for it to heat up). This would normally only occur if the 5V line rises above 5.5V and the *USB Port Protector* is shunting current in order to prevent it rising further.

In fact, the *USB Port Protector* does very little as long as the USB supply voltage is in the normal range of 0V to 5.25V and the D– and D+ lines are in the normal range of 0-3.3V. Green LED1 lights up to indicate power is present, but that's about it. The *USB Port Protector* draws around 3mA in this condition.

If the 5V rail is pulled negative, ie, below 0V (eg, you've accidentally shorted it to the output of a transformer or some other supply rail) then schottky diode D3 will conduct. This prevents V_{CC} from going below about –0.5V.

D3 is a high-current diode, capable of handling 15A continuously and 275A for around 5ms, so it makes a very effective clamp. It limits the voltage on V_{CC} to –0.55V at 15A, so your PC is safe from damage from negative voltages on the supply line.

Should the overload condition persist, either PTC1 will limit the overload current to a safe level or F1 will blow, disconnecting the compromised circuitry from your computer.

Clamping positive voltages

It's even more likely that you might accidentally short the 5V rail to a higher voltage; eg, 12V from a car battery. Just think of the heavy currents which will fry anything connected to it! The *USB Port Protector* has active and passive systems to handle this situation.

The active system is the first line of defence. It comprises high-current PNP transistor Q1 and shunt voltage reference REF1. The 1.2kΩ/1kΩ resistive divider across the 5V supply feeds 45.45% of the supply voltage to the adjust terminal of REF1. It sinks current into its cathode terminal as soon as this adjust terminal exceeds +2.5V.

So, given the voltage divider, that means that it will sink current when the supply exceeds 5.5V (2.5V ÷ 45.45%). This will cause a voltage to develop across the 470Ω resistor and once that voltage exceeds around 0.7V (Q1's base-emitter voltage), Q1 will switch on and shunt the 5V supply rail, pulling it down.

In this manner, REF1 and Q1 act to limit the 5V supply rail to just over 5.5V. Q1 is capable of handling more than 10A, but since there will be 5.5V between its collector and emitter, it can only do that for a very short time before it overheats. But at the same time PTC1 will rapidly heat up and increase its resistance, to limit that current. And in any case, if the current exceeds 3A, for example, the fuse will very quickly blow before Q1 is damaged.

So REF1/Q1 act together as a very precise and very fast clamp. When REF1 is sinking current from Q1's base, Q2 will also normally switch on as its base is also pulled around 0.7V below its emitter, via the 10kΩ resistor. This will light up red LED2, indicating that the clamp is operating and that you have a problem. LED2's current is limited by its low base current and relatively fixed gain (h_{FE}).

REF1 can sink up to at least 100mA and Q1 has a current gain (h_{FE}) in the hundreds, so Q1 is more than capable of passing its full peak current rating of 24A in this circuit.

Note that LED2 may go out if there is a persistent overload, since when Q1 heats up, its base-emitter voltage will drop and it may drop low enough below Q2's base-emitter switch-on voltage that it will no longer switch on. But chances are that PTC1 and/or F1 will have acted to limit the fault current by that stage anyway.

The only problem with the clamp provided by Q1 and REF1 is the reaction time. It takes a short time for REF1 to react to an increase in the feedback voltage and it also takes time for Q1 to switch on – around a microsecond.

Passive clamping

This is why we also have a transient voltage suppressor (TVS1) connected across the 5V supply rail. It's a passive

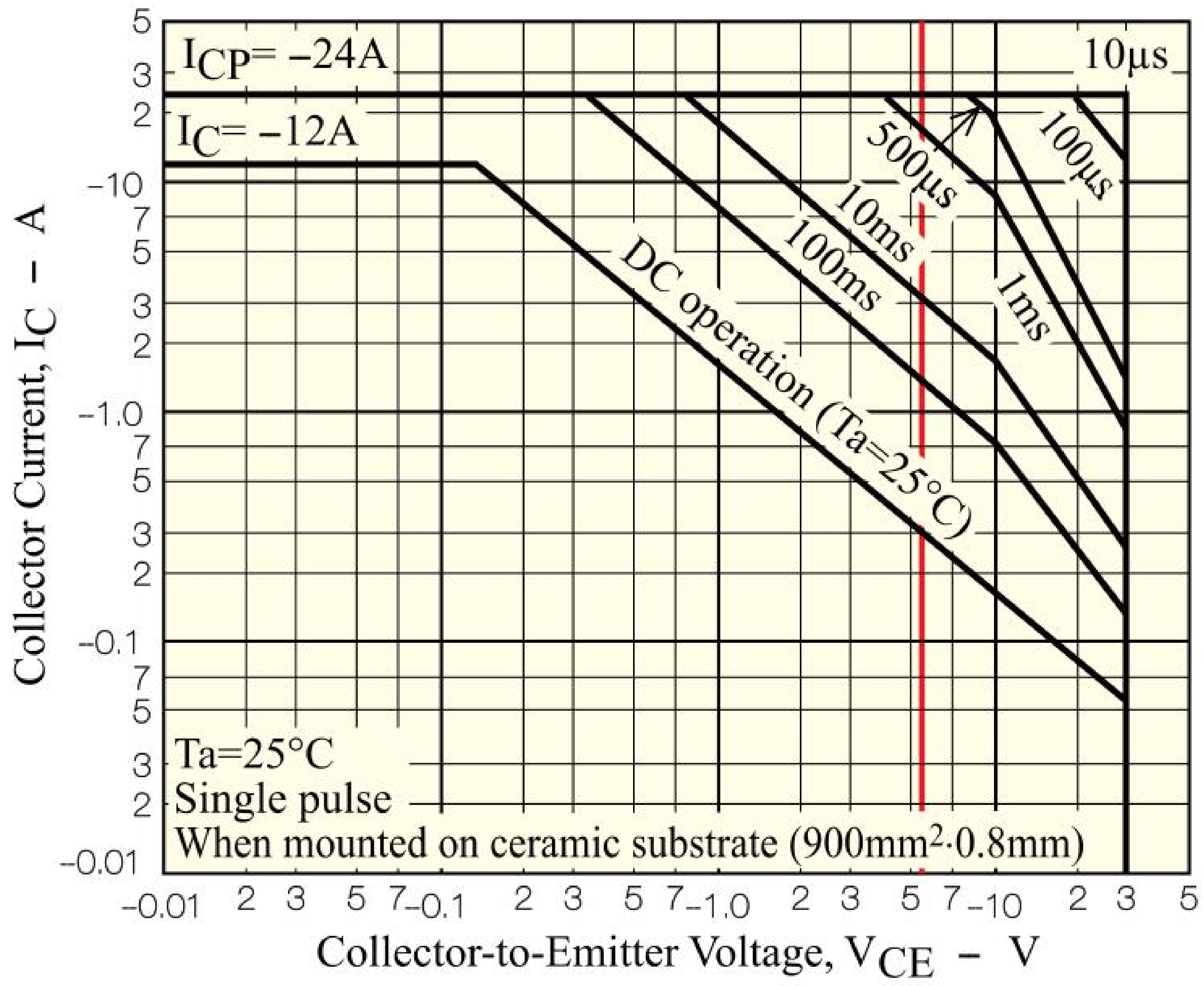


Fig.2: safe operating area (curves) for the ECH8102 PNP transistor, used in this device as a protective shunt. The vertical red line corresponds to a shunt voltage of 5.5V and its intersection with the SOA curves shows how long the transistor is guaranteed to survive at various collector current levels.

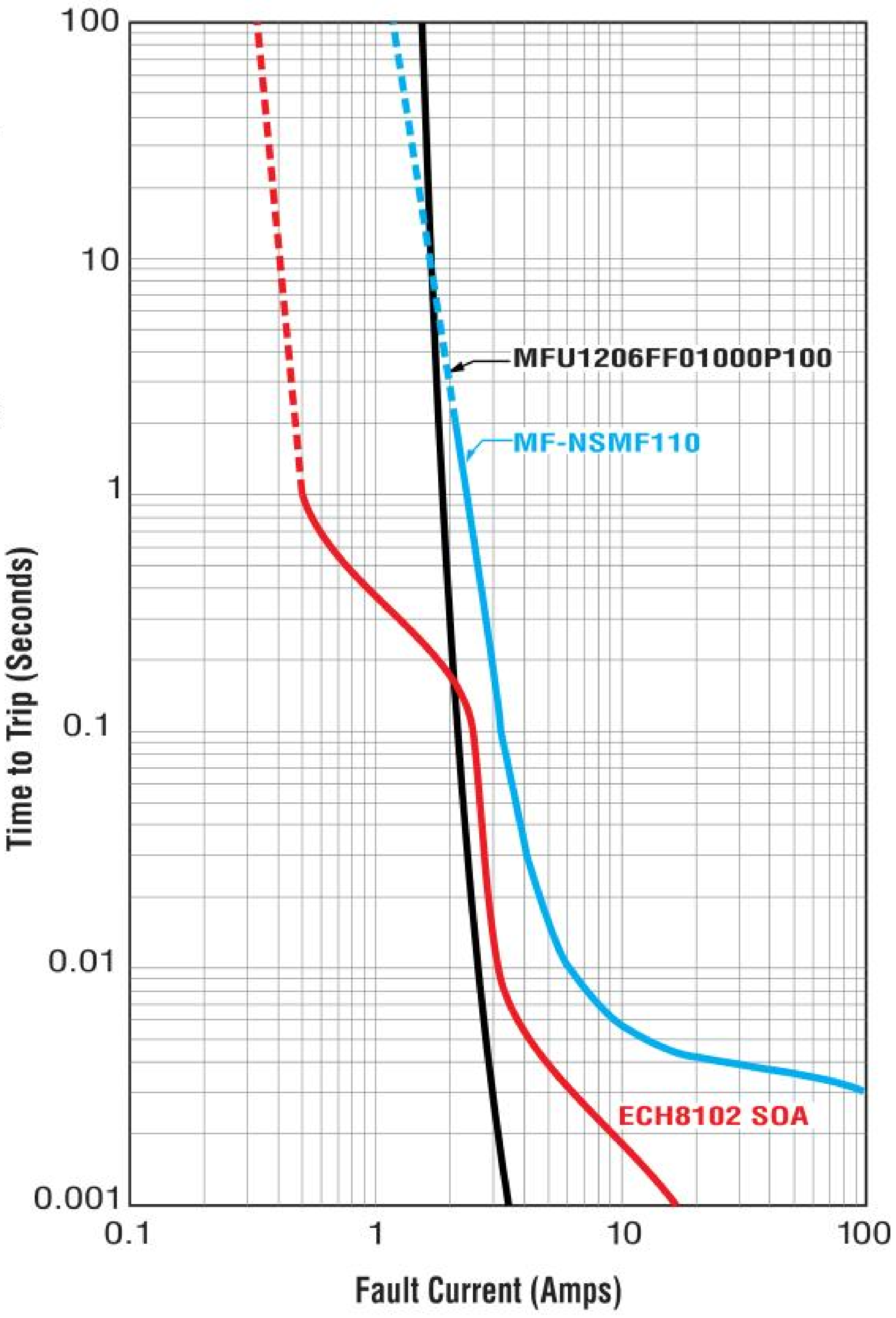


Fig.3: the fuse blow time for F1 (black) and 'trip' time for PTC1 (blue) at various current levels. The relevant portion of Q1's SOA curve from Fig.2 is plotted in red and you can see that F1 will protect Q1 for fault currents above 2A.

device which will react more-or-less instantly to excessive voltage.

But like most zener-type devices, the difference between the voltage at which it will start to conduct current and the voltage across it when a large current is flowing is quite large. We've selected the most suitable device possible, but it's still not ideal.

The 'working voltage' for TVS1 is defined as 5V but it's designed to pass only 1mA or so at 6.0V. The clamping voltages are specified as 9.8V at 1A and 13.5V at 42A.

So clearly, we can't rely on this device to protect the PC since it would allow quite a high voltage to be fed back in before taking effect. Hence our dual-action strategy, with TVS1 there to limit very brief, high-voltage excursions (eg, a static discharge) and also to 'fill in the gaps' for the short period until Q1/REF1 are able to switch on and shunt the fault current.

Protection for the signal lines

We've also included 3V transient voltage suppressor TVS2 and dual schottky diodes D1 and D2 to protect against damaging voltages being fed in via the D+ and D– signal wires. This is unlikely, since these lines normally go straight to some sort of USB/serial adaptor or micro on a development board, and so there aren't many exposed components to accidentally short.

But it's still possible that a high voltage fed into your +5V rail (or +3.3V rail, or some other supply point) could damage the USB/serial adaptor or microcontroller and allow current to flow through into the D+ and/or D– lines. So we decided that we should provide at least some protection for these lines, as well.

The half of dual diodes D1/D2 that connects between ground and the signal line prevents them from being pulled too far below ground.

We're using smaller diodes here since a large diode would have too much capacitance and would interfere with USB signalling. But these diodes are still rated at 300mA continuous and 1.25A for 10ms, with a forward voltage below 1V up to several hundred milliamps. So they should provide decent protection.

TVS2 has a breakdown voltage of around 3.6V at 1mA and a clamping voltage of 6.5V at 25A. So the combination of D1/D2 and TVS2 should conduct significant current away from the D+/D– lines well before their voltages reach 5V. Most USB ports would not be damaged by these voltages.

We can't put a voltage suppressor like TVS2 directly between the D+ and D– lines and ground because it would have far too much capacitance. But the series diodes between D+/D– and TVS2 have a much lower capacitance that's effectively in series with that of TVS2, so they have virtually no effect on signalling. We tested our prototype with a 'hi-speed' USB card reader and it functioned normally.

Is it bulletproof?

In a word, no, but if it does fail, the *USB Port Protector* is likely to fail in such a way that it still protects your computer. While our testing shows that it's robust and can handle significant overloads without damage, if you apply just the right (worst possible) combination of voltage and current, it may be possible to blow Q1 or TVS1 before fuse F1 blows.

Still, our testing suggests that the most likely outcome of a serious overload is for F1 to blow and at least it's cheap and (relatively) easy to replace.

The difficulty in designing a circuit like this to be able to withstand anything you can throw at it is that in order to effectively protect against a high current source being connected to the V_{CC} line, it needs to absorb quite a lot of power in a brief period. And while the PTC and/or fuse should ideally cut the power to protect the other components, they may not be fast enough.

Fig.2 shows the 'safe operating area' (SOA) curves for transistor Q1, taken from the ECH8102 data sheet. We've added a vertical red line to show the typical voltage of about 5.5V across Q1 while it is conducting.

While this is a high-current transistor, it is quite tiny, so if a high current is applied, it will quickly overheat and might fail. As shown in Fig.2, it's guaranteed to survive 24A at 5.5V (132W!) for somewhere between 500µs and 1ms. For longer periods, the maximum allowable current is lower; around 3A (16.5W) for 10ms, 1.5A (8.25W) for 100ms and 300mA (1.65W) continuously.

Beyond this it may survive – but that isn't guaranteed. Our testing has shown that for a single pulse, these ratings are very conservative. But it's good practice to design a circuit to stay within these ratings.

The 'trip' times for PTC1 (blue) and F1 (black) are shown in Fig.3. We've also plotted the relevant portion of the SOA curve for Q1 in red so that you can compare them. As you can see, F1 responds considerably faster than PTC1, and in fact is very likely to blow before Q1's SOA is exceeded for currents above 2A.

For fault currents between 300mA and 2A, it's possible that Q1 will overheat and fail before either F1 blows or PTC1 acts to limit the current. And in fact, PTC1 is not guaranteed to do anything for fault currents below 1A. You will need to notice red LED2 lighting and resolve the fault yourself.

Still, as we said above, the ratings for Q1 seem to be pretty conservative and as long as the overload is limited to no more than a second or two, we would expect it to survive.

Parts list
USB Port Protector

1 double-sided PCB, coded 07105181, 32.5 x 19mm
1 PCB-mount USB Type A horizontal plug (CON1)
1 PCB-mount USB Type A horizontal socket (CON2) [eg, Altronics P1300]
1 SMD fuse, 3216/1206 package, 1A super fast blow [Vishay MFU1206FF01000P100]
1 SMD 1.1A PTC thermistor, 3216/1206 package [Bourns MF-NSMF110-2]
1 30mm length of 20mm diameter clear heatshrink tubing

Semiconductors
1 AN431AN shunt reference IC, SOT-23 (REF1)
1 ECH8102 12A PNP transistor, ECH8 (Q1)
1 BC856 100mA PNP transistor, SOT-23 (Q2)
1 high-brightness green LED, 3216/1206 package (LED1)
1 high-brightness red LED, 3216/1206 package (LED2)
1 CDSOD323-T05S transient voltage suppressor, SOD-323 (TVS1)
1 SM2T3V3A transient voltage suppressor, DO-216AA (TVS2)
2 BAT54SFILM dual 300mA schottky diodes, SOT-23 (D1,D2)
1 15A 30V schottky diode, DO-214AB (D3; MCC SK153)

Capacitors
1 100nF SMD X7R ceramic, 3216/1206 package

Resistors (all SMD 3216/1206 package, 1%)

1 47kΩ	(coded 4702 or 473)
1 10kΩ	(coded 1002 or 103)
1 1.2kΩ	(coded 122)
1 1kΩ	(coded 102)
1 470Ω	(coded 471)

Looking at Fig.2, you may wonder why we've bothered with the PTC at all, given that its 'trip' current is higher than the fuse blow current over most of the graph. But keep in mind that PTC1 is considered to be 'tripped' when it has reached a high enough resistance value to keep the fault current below 2.2A.

It will still have some effect in reducing the fault current, even at lower current levels and shorter time spans, because its resistance will start to increase well before it has fully tripped.

And you also have the option of replacing F1 with a zero-ohm resistor (or just soldering across the pads) and relying on PTC1 to limit fault currents. This does increase the risk of blowing Q1 in a serious fault (although, as we said, it's pretty robust) but doing so would also increase the chance that the unit will survive a moderate overload unscathed and you won't have a blown fuse to replace.

Note that while replacing Q1 is a bit of a pain, it's actually quite cheap (under 50p) so if Q1 does 'throw itself on the grenade' and fail while protecting your computer from damage, at least it isn't an expensive failure.

Sourcing the parts

Most of the parts are surface-mount devices (SMDs) and they are all available from Digi-Key, Mouser or element14.

Construction

The *USB Port Protector* is built on a double-sided PCB available from the *PE PCB Service*. It measures 32.5 x 19mm and is coded 07105181. All but four of the components are mounted on the top side of the board, as shown in the overlay diagrams, Figs.4 and 5 and matching photos. The only through-hole components are the USB plug and socket. (By the way, you may notice a minor difference between the overlay diagrams and the PCB photos: we've changed TVS2 to a more suitable part since building the prototype.)

Most of the parts are fairly easy to solder, although some of them are quite close together, to keep the unit compact. It's easiest to do in the following order.

Start with transistor Q1. This is in a fairly small ECH8 package, with four short leads on each side. The good news is that most of the adjacent leads are connected together so it doesn't matter if you bridge the pins when soldering (in fact, it's pretty much unavoidable). Pin 4 is the base connection and you need to make sure it doesn't short to pin 3, the emitter.

Start by identifying pin 1. There is a dot printed in the corner on the top of the package, but you will need a magnifier and good light to see it. Orient the part so that it matches the pin 1 markings on the PCB and smear a thin layer of flux paste on all eight of its pads.

Apply a tiny amount of solder to the pad for pin 4, then heat this solder while sliding the part into place. Check that the other seven pins are correctly located above their pads using a magnifier. If not, re-heat the solder joint and carefully nudge the part. Repeat as necessary until it's lined up, then solder the four pins on the opposite side of the package. These are all joined together so you can do it as one big solder joint.

Now apply solder to the three remaining pins and add a bit of fresh solder to pin 4 as well. To tidy up the solder joints, apply a little more flux paste on top of the solder and then use some solder wick to remove the excess. Clean up the flux residue with some methylated spirits, isopropyl alcohol or other flux cleaner and then inspect it visually to ensure all the solder joints are good.

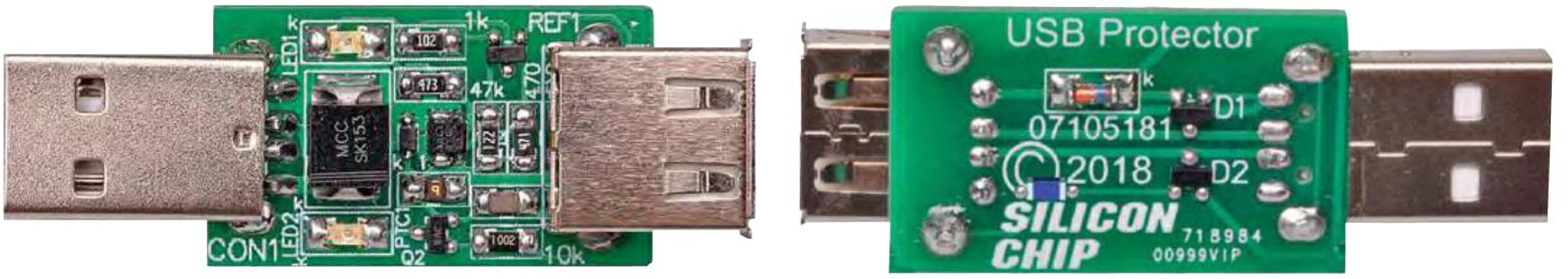
That's the trickiest part out of the way. Next, solder TVS1 in place, next to Q1. It's fairly small and its cathode stripe will not be terribly obvious; so again, use magnification to identify the cathode and orient it correctly before tacking it in place and soldering the opposite pin.

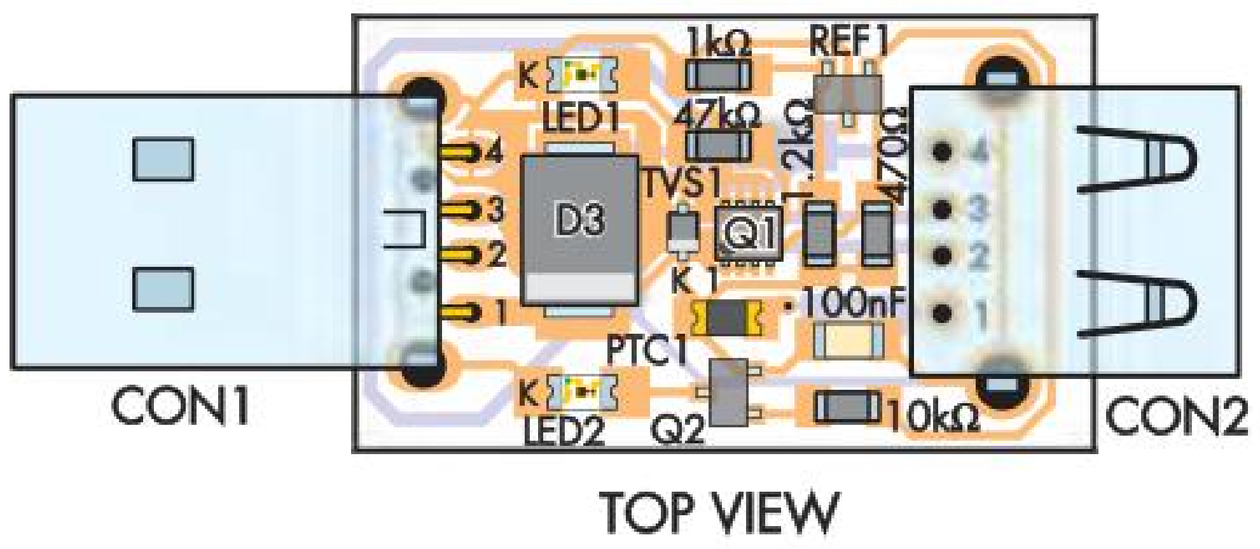
Now solder the SMD passive components in place; this includes five resistors, one capacitor and the PTC thermistor. None are polarised; just be careful to fit each in the location shown in Fig.4. The resistors will be printed with a small code indicating their value (eg, 1.2kΩ code is 122; or 12Ω x 10^2) but the capacitor will not be marked.

The resistor codes are also shown in the parts list opposite.

The next components to mount are reference REF1 and transistor Q2. These are in identical SOT-23 packages so don't get them mixed up after taking them out of their packaging. They are polarised, but have three pins each, so the orientation is obvious – see the pinouts in Fig.1.

Next are the two LEDs. Usually, the cathode is marked with a green dot but sometimes the anode is marked instead. The easiest way to check is with a DMM set on diode test mode. The LED will light up with the red probe connected to the anode and black to the cathode. You can confirm the colour at the same time. Note that some DMMs (eg, those powered by two AA cells) may not apply sufficient voltage to light up a green LED.





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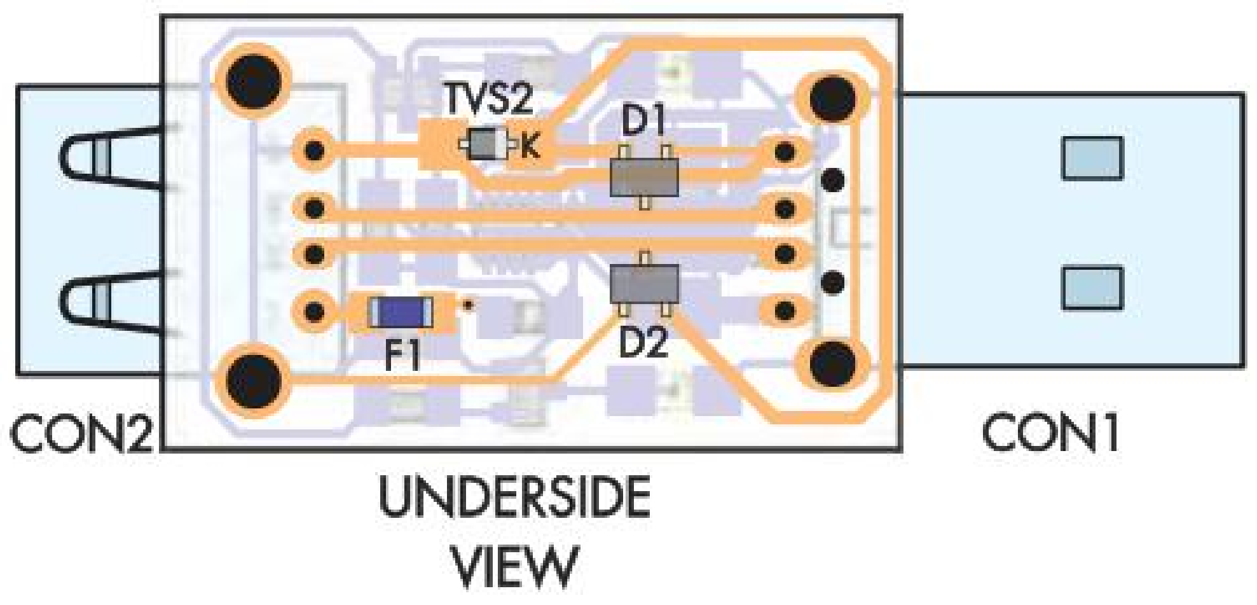


Figs.4 and 5: top and bottom overlay diagrams for the *USB Port Protector*. Use these as a guide during construction. Be careful with the polarity for TVS1, TVS2, Q1 and LEDs1 and 2. It's easiest to start by fitting Q1 and TVS1, then the remainder of top-side SMD components, then the bottom-side components and finally, CON1 and CON2. The matching photographs above are reproduced close to twice actual size, for clarity.

Solder these where shown on the overlay diagram; LED1 is green while LED2 is red and the cathodes are oriented towards the USB plug, as shown by the 'K' markings on the PCB.

Now solder schottky diode D3 in place. Add a little flux paste to the pads first as it's quite large, but the procedure is much the same as for the other two-pin devices. Just make sure you apply the iron for long enough to form good solder fillets between the PCB and terminals of the device.

Then flip the board over and fit the four remaining SMDs on the bottom side, as shown in Fig.5. D1, D2 and ZD1 are polarised; also pay particular attention to the location of the cathode stripe on ZD1. The fuse is not polarised.

Finally, fit the USB plug and socket as shown. Both need to be pushed down firmly onto the PCB before soldering. The plug has a notch on the underside which the edge of the PCB fits into. Note that the USB plug pins may be quite short and may not protrude very far through the bottom of the PCB, so it's a good idea to solder them on both sides. Just make sure you don't accidentally bridge the pins.

TVS2 note

The TVS2 (SM2T3V3A) is in an unusual package – its anode runs the full length of its body. When soldering it, ensure that its body is not on top of the cathode pad (marked K) or else it could short out the USB data signal lines.

Testing

Inspect the board to verify that all the solder joints are good and that you have no unwanted bridges, then plug it into a USB port on your PC. If you have a USB charger, you could use that instead. Check that the green LED lights up but the red LED should not.

You can then carefully measure the voltage across D3. You should get a reading in the range of 4.5-5.25V (usually quite close to 5V), with the red probe to its cathode (striped) end.

Now plug a small device like a USB card reader or flash drive into the socket and verify that it powers up correctly. Try reading the contents of the card/flash drive on your PC and verify that it works normally without any unexpected disconnection events.

If you want to verify that the *USB Port Protector* will definitely protect your computer, you will need a ~6V supply and a resistor with a value between 2.2Ω and 10Ω.

Unplug the *USB Port Protector* and anything that's plugged into it and use a clip lead to connect the USB socket shell to the ground terminal of your 6V supply. Connect one end of the test resistor to the positive output of the 6V supply (battery pack, plugpack...) and then touch the other end of the resistor to the USB socket pin that's immediately adjacent to fuse F1, on the underside of the board.

If you can do this while looking at the top of the board, you should see both LED1 and LED2 light up. LED2 indicates that the protection is operating. If you have a helper, they could measure the voltage across D3. It should be close to 5.5V. This confirms that the device is working.

Using it

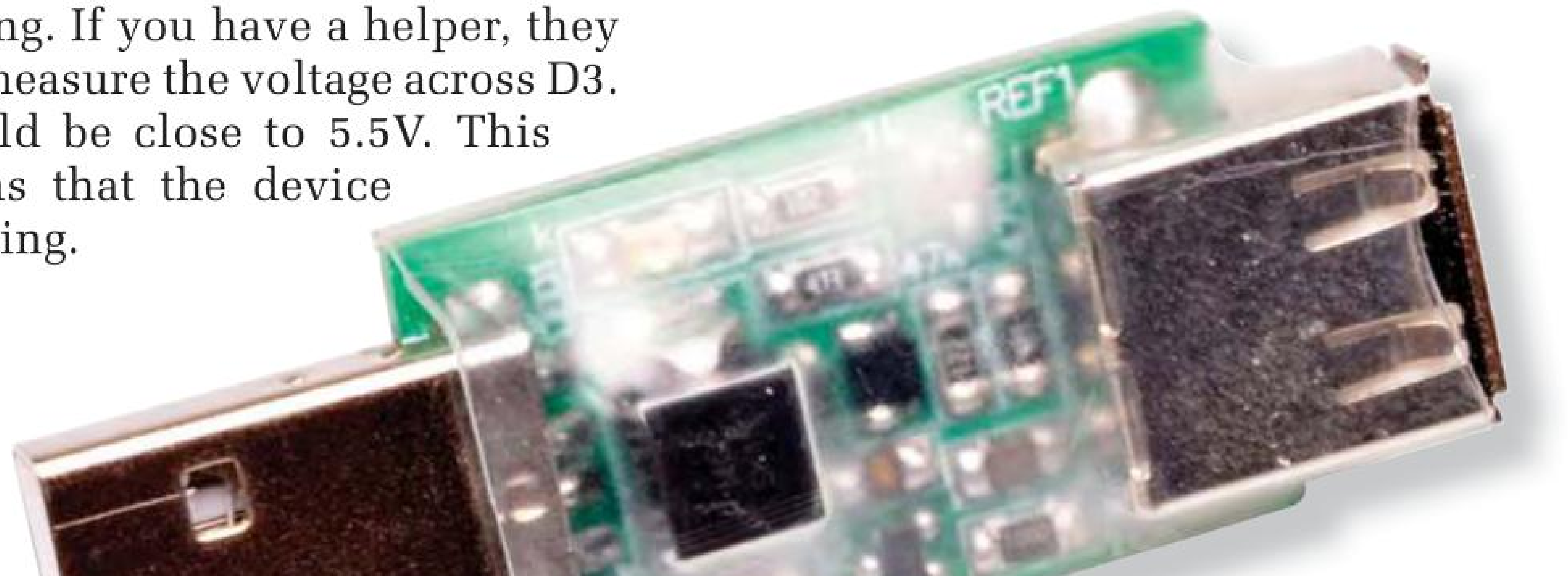
To avoid accidentally shorting the 5V supply or either of the signal lines during use, we suggest you encapsulate the entire device in a short piece of heatshrink tubing, as shown above. Clear tubing is convenient since you can still see the components – but any colour will work.

Cut the tubing so that it covers the entire USB socket, up to the lip that's around the open end, and the base of the USB plug, up to where it projects from the PCB. Then it's just a matter of applying a little heat, eg, from a hot air gun or hair drier.

Rotate the assembly until the tubing has shrunk into place and try to avoid burning yourself in the process. If it gets too hot to hold, put it down and let it cool before shrinking the remainder of the tubing.

If you manage to blow the fuse, you will simply have to cut the tubing off, desolder the fuse, clean the old solder off using flux paste and some solder wick, solder a new fuse in place and apply a fresh length of heatshrink tubing.

Or if you're really clever, you may be able to cut a flap in the tubing around the fuse, replace it and then glue the flap back in place.



We finished our *USB Port Protector* with clear heatshrink tube... just in case A.P. managed to drop something into the PCB!

MIDI Ultimate Synthesiser

Soundtronics
MFOS

Part 4 | by Paul Cooper | soundtronics.co.uk

For Part 4 of the MIDI Ultimate Synthesiser build, we are going to add the circuits for the modulation sources. Control voltages can also enact many of the manual potentiometer controls used to modify the characteristics of the sound being produced. These control voltages take various forms, including cyclic waveforms, random voltage sequences and envelope generators. Envelope generators can be single shot or continually re-triggered to make more complex cyclic waveforms.

We briefly discussed each of the modulation sections of the *MIDI Ultimate* in *Part 1*. The patching diagram (Fig.5, *Part 1*) shows how the modulation sources can be routed to the oscillators (VCOs), filter (VCF) and the voltage-controlled amplifier (VCA). Sounds that have a constant frequency, amplitude and harmonic content can be rather bland. Being able to change some or all of these characteristics over time using modulation sources is what brings the synth to life.

No synth will be complete without two main modulations sources, the envelope generator (ADSR) and the low-frequency oscillator (LFO). Recognising how important these are, the *MIDI Ultimate* includes two of each of them. Another modulation source at our disposal is the sample and hold (S&H) generator that outputs a snap shot or sample of a waveform periodically. The waveform being sampled by the S&H can be any of the wave shapes from LFO 2 or from the noise source we built in *Part 3*. The rate of sampling can be manually adjusted, as can the glide, or *portamento*, from one note to the next.

Those familiar with analogue synthesis can skip this next section on modulation, but for those new to analogue synths the following will aid in understanding the synthesiser's operation and use.

Modulation

The waveforms from the LFOs can be patched to the VCOs to vary their frequency – this is know as 'vibrato'. As the LFOs have four choices of wave shape, the *vibrato* effects can be quite different. A square wave will cause the VCO to alternate between two frequencies, whereas a triangle wave shape will cause the VCO frequency to rise and fall linearly between two frequencies. The difference between the two frequencies is determined by the amount of modulation depth set on the VCO(s). The starting frequency can vary with what key is being pressed on the MIDI keyboard or other modulation sources.

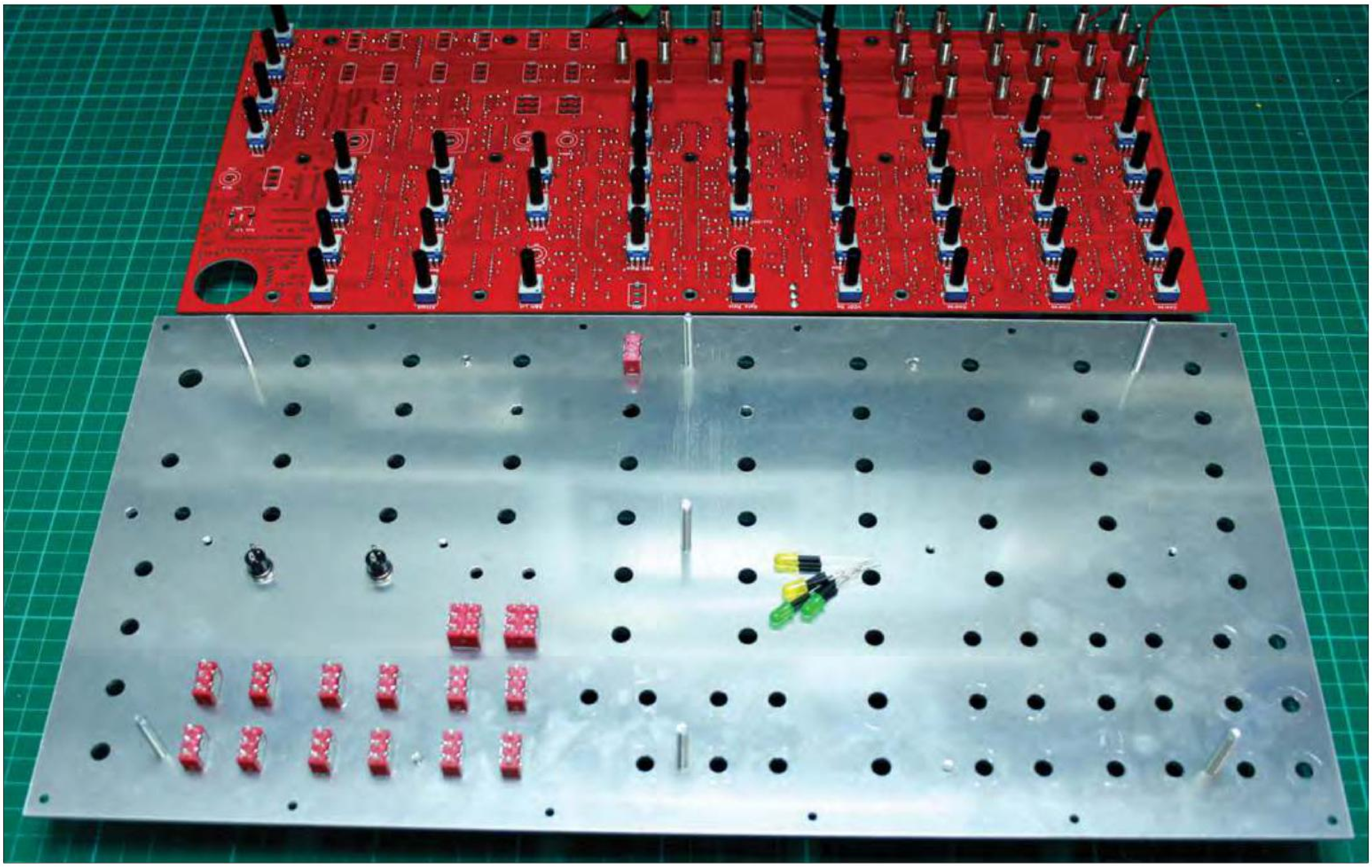


Fig.59. Switches mounted into their corresponding positions, pots clipped into position and LEDs ready for inserting into the PCB.

The LFOs can also modulate the VCA to create a varying amplitude (volume) of the sound, better known as 'tremolo'. Some manufacturers of guitars, for example, have incorrectly called the vibrato arm or bar a 'tremolo arm'. So, to be clear:

- *Vibrato* is a change in *pitch*
- *Tremolo* is a change in *volume*.

Finally, the LFOs can be used to modulate the cut-off frequency of the VCF and thus alter the timbre of the sound between two frequency cut-off points.

The envelope generators can be used in AR or ADSR mode (see Fig.20 to Fig.22 in *Part 1*). In ADSR gate mode, on a gate pulse, the envelope generates a voltage that initially climbs to the maximum voltage level over a period of time called the attack period (A). When the maximum voltage has been reached, the voltage will reduce to a voltage level set by the sustain (S) potentiometer in the decay (D) period. The output will now stay at the sustain voltage until the gate pulse is removed and the output voltage decays to zero over the release (R) period. This is why the envelope generator is frequently just called an ADSR. Variations on the ADSR – depending on mode and potentiometer settings – can also achieve an envelope just consisting of the AR stages. Envelope generators can modulate the VCOs (ADSR2 only), VCA and VCF just like the LFOs, although they are not often used for VCOs. Their principle function is with the VCA for amplitude modulation; imagine striking a piano key where you would hear the note immediately but it would take time for the note to diminish. Many small synths like the *MIDI Ultimate* only include one AR let alone an ADSR, so why do we have two ADSRs? Having two ADSRs means you can simultaneously control both the VCA and the VCF with different envelopes, offering a far greater range of sounds and enables the user to better emulate what happens with traditional instruments. Triggering an ADSR can be from a MIDI key being pressed or from the dedicated repeat gate generator, or even from the S&H gate generator.

Last but not least is the S&H; feed in a sawtooth waveform from LFO2 and the synth generates a staircase of increasing voltage levels. This control voltage can not only be used to modulate the VCOs for a sequencer effect, but also creates excellent timbre effects when used with the VCF. Select other source waveforms for different effects, including the white noise source for random voltage levels.

Construction

Techniques used for this month's construction are similar to that in *Part 2* and *Part 3*, so do refer back to the previous articles as needed. Additionally, two pushbuttons and four LEDs are mounted on the pot/switch side of the PCB. We now have six sections on the main PCB to populate with components, these are:

- ADSR1 and ADSR2
- LFO1 and LFO2
- Repeat gate generator
- S&H

I recommend soldering in the components for all six sections before moving onto the switches, pots and LEDs (see Fig.59). As before, do not insert the chips until you are ready to test, and then only one section at a time. Testing is covered at the end of this article.

The LEDs require spacing off the PCB by 8mm, we achieved this by cutting some sleeving and sliding over each leg

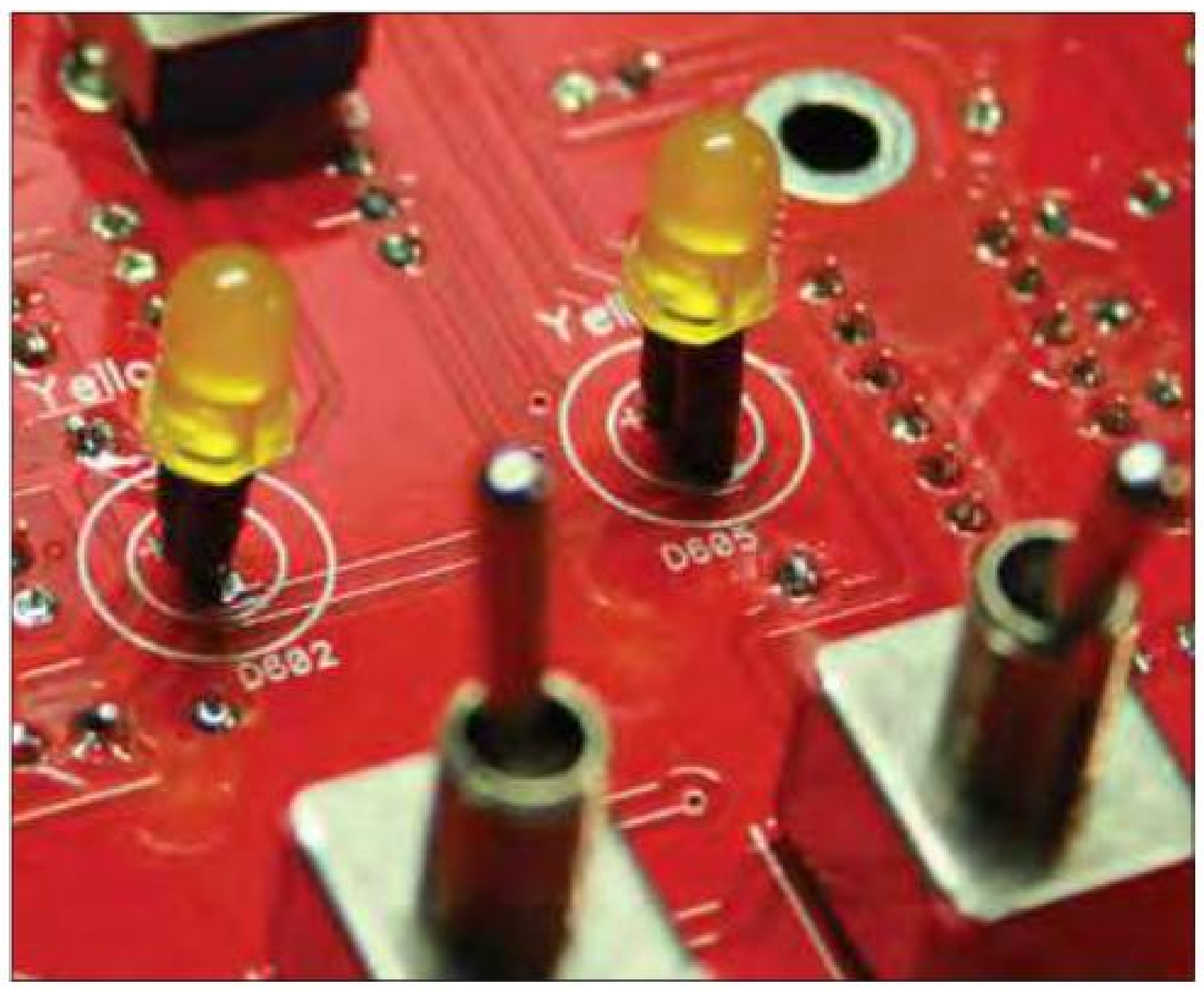


Fig.60. Finish height of LEDs is 16mm above the PCB.

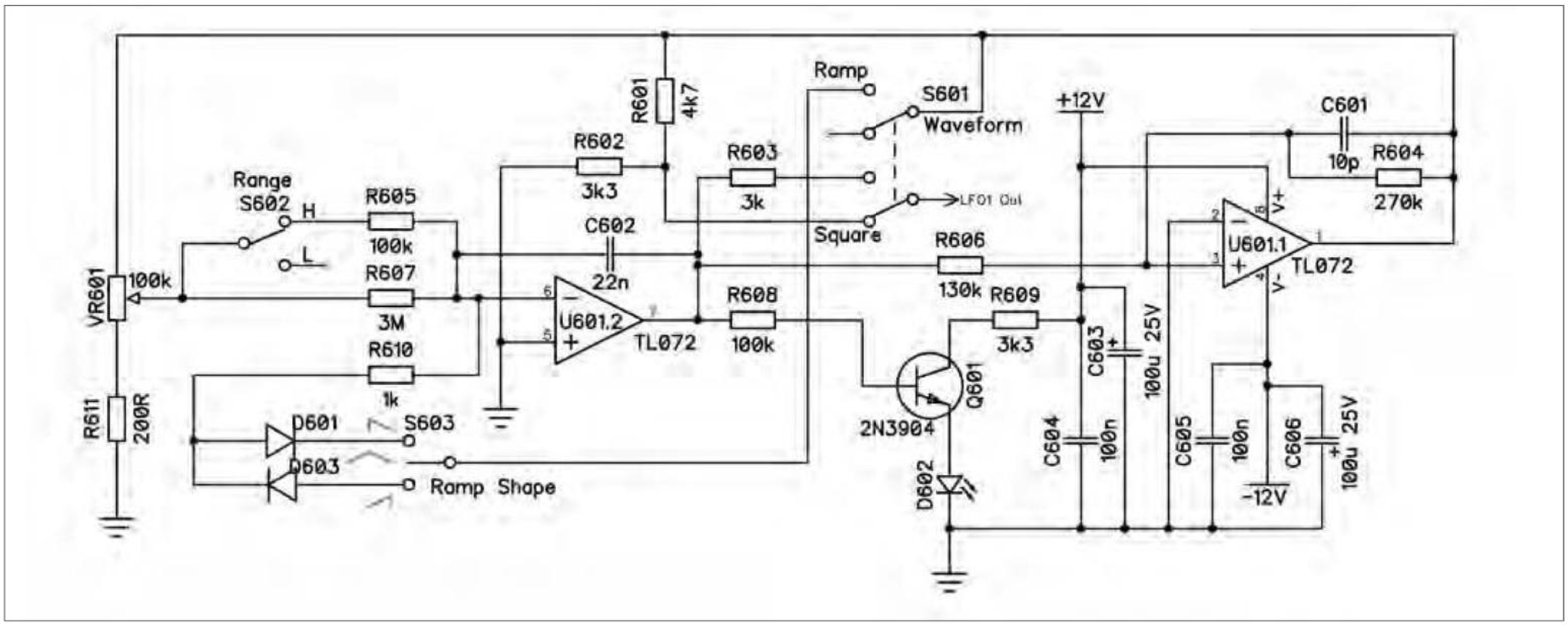


Fig.61. *MIDI Ultimate* LFO1 circuit.

of the LEDs. First check that the LEDs go into the holes in the aluminium panel as we have had some reports that the hole diameter may be a bit tight. If necessary, run a 5mm drill through the LED holes to make sure and deburr afterwards. The long leg of the LED goes into the pad marked with a '+'. After soldering the LEDs, do not clip their legs just yet in case you need to adjust their height. The finish height of the LED is around 16mm above the PCB (see Fig.60).

LFO circuit description

The two LFOs are identical in operation, so only LFO1 will be described here. The LFO has two frequency ranges selected by S602. 'High' gives a frequency range of a few Hz to about 300Hz, 'Low' from 0.1Hz to 30Hz. VR601 sets the frequency within the range selected. Two switches are used to select the output waveform, S601 selects 'Ramp' or 'Square', if ramp is selected, then S603 further selects what type of ramp waveform, 'Saw', 'Inverted Saw' or 'Triangular.

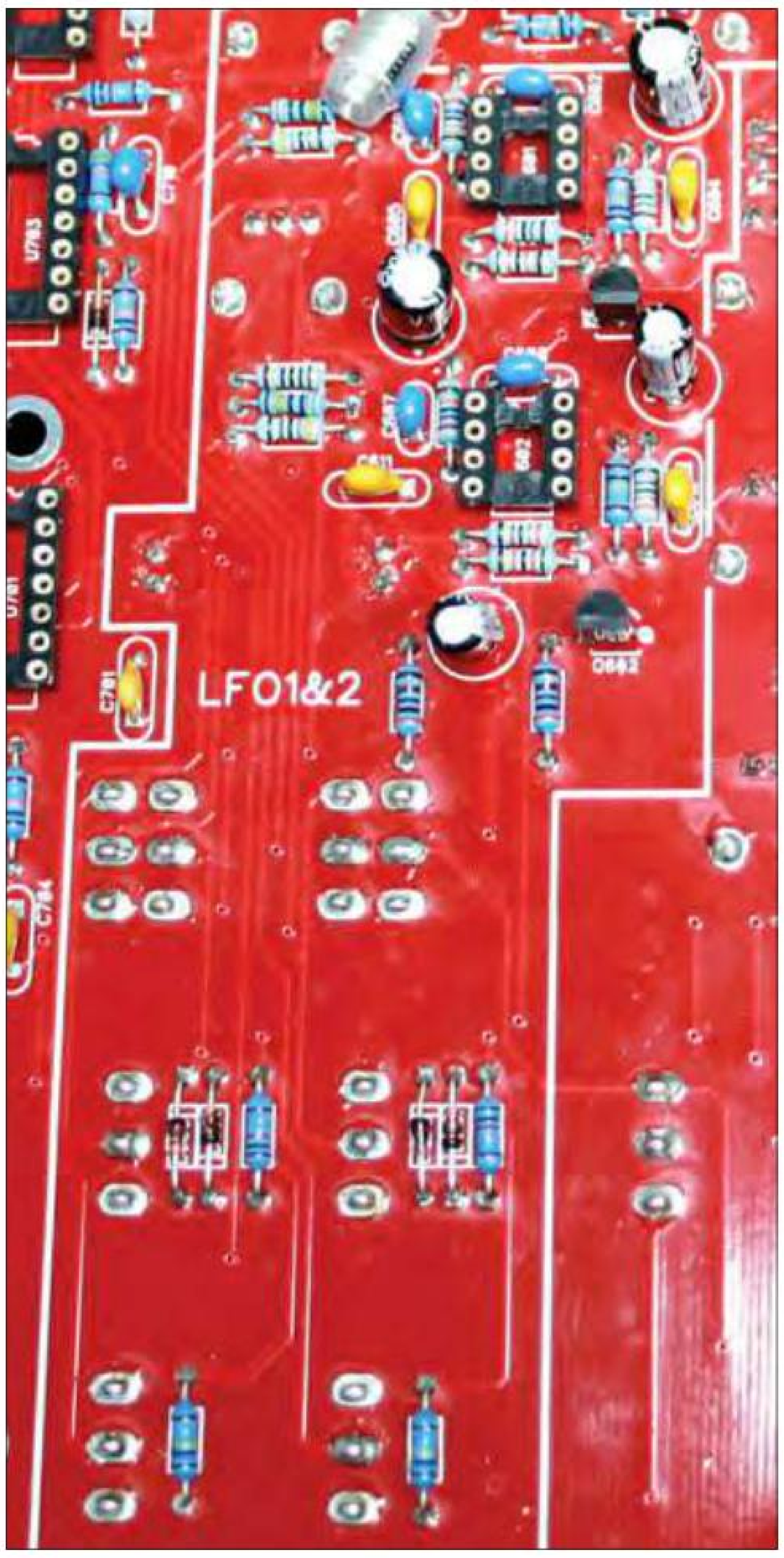


Fig.62. LFO1 and LFO2 assembled.

This LFO integrator/comparator design is not unique; Jake Rothman described a similar circuit he used in his *Gen X-1 Stylephone* (see: *EPE*, August 2018, Fig.13) and again in his article from October 2018 (see Fig 61).

Referring to the LFO1 circuit diagram in Fig.61, U601.2 forms an integrator and U601.1 a comparator. The output of the comparator U601.1 feeds back into the integrator U601.2 via the frequency pot VR601 and R607 or R607 in parallel with R605 if the range switch has 'High' selected. The setting of the potential divider VR601 determines the current flow into/from integrator U601.2 and how quickly the integrator capacitor (C602) is charged/discharged, and thus the rate at which the output of the integrator ramps up and down. This ramping voltage is fed into the comparator's non-inverting input; the inverting input is connected to ground. The comparator has significant hysteresis due to feedback resistor R604.

Oscillation

When the comparator output is high at 11V, the integrator capacitor (C602) is charged up via VR601 and the integrator output voltage ramps down towards –11V (inverting). When this voltage falls below about –5V, the comparator output switches to –11V. Now the integrator capacitor is discharged via VR601 and the integrator output begins ramping up towards +11V. When it reaches around +5V, the comparator output switches to +11V and the process repeats, resulting in oscillation.

Square operation

When 'Square' is selected on S601, the LFO waveform is the output of the comparator reduced to ±5V by potential divider R601/R602, which is toggling between high and low and thus produces a square waveform.

Ramp operation

When 'Ramp' is selected on S601 and 'Triangle' on S603, the output via R603 is now from the integrator output which is as described above, a ramp up and ramp down waveform already at ±5V.

Sawtooth Ramp operation

To generate the sawtooth and ramp (inverted sawtooth) waveforms, diode feedback from either D601 or D603 and R610 is introduced. This feedback has the effect of cutting short either the rising or falling ramp to around 1ms, depending on the waveform selected. So, when S603 is selected to 'Sawtooth' and with the comparator output low, the integrator ramps high in 1ms through D601 which in turn forces the comparator output to go high. With the comparator output high, D601 is reversed biased so the normal ramping through VR601 occurs, as described above. When 'Ramp' is selected, D603 is in place of D601 in the feedback loop. As D603 is reversed in relation to D601, you get the opposite effect to the sawtooth, which is a slow ramp up, and a 1ms ramp down. This is a fairly simple circuit, which has one side effect, when either sawtooth or ramp is selected, the LFO frequency is doubled because one half of the ramping cycle is cut short to 1ms. An LED, D602 driven by Q601 illuminates in sync with the output waveform.

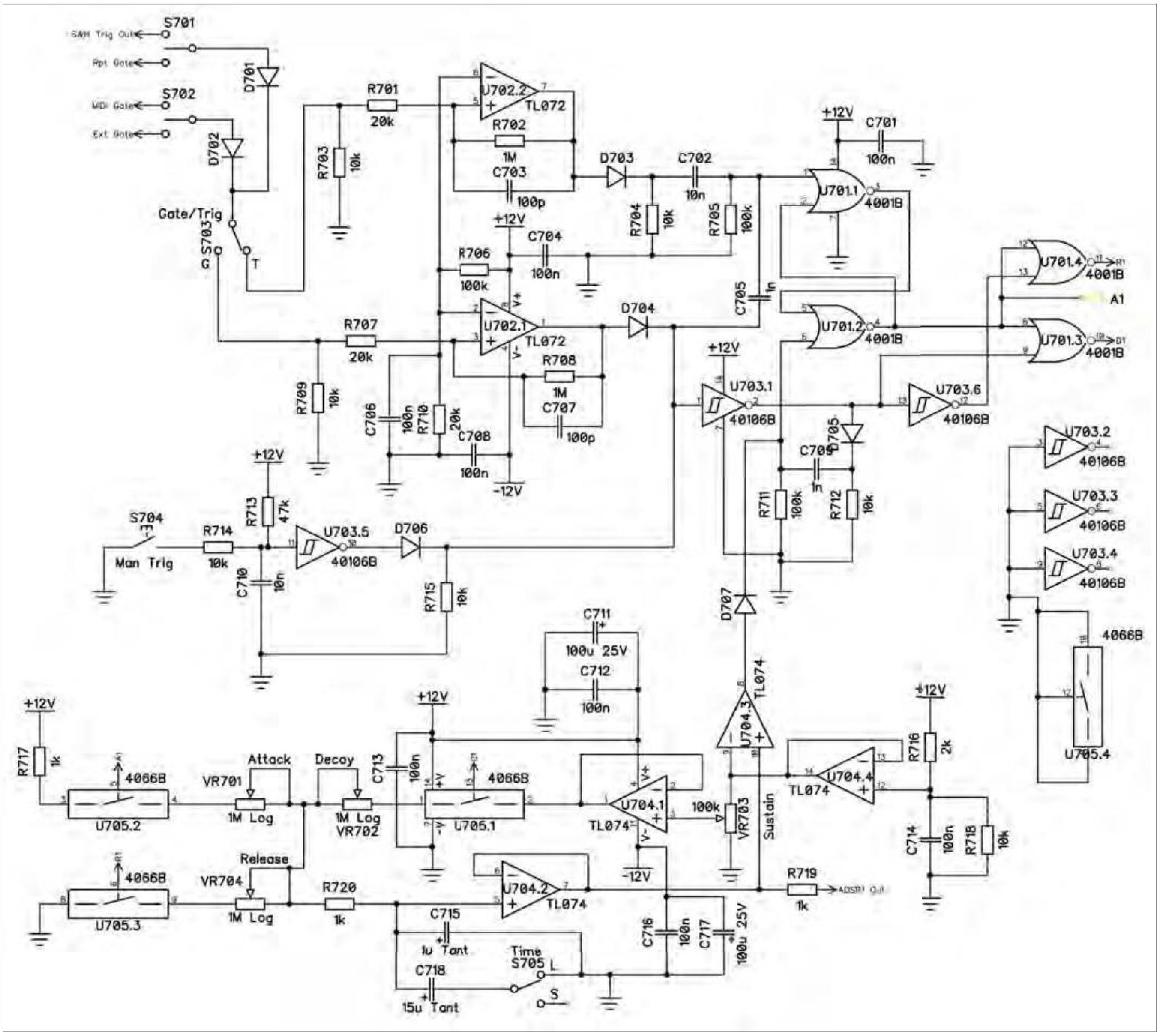

S701
S&H Trig Out
Rpt Gate
S702
MIDI Gate
Ext Gate
D701
D702
Gate/Trig
S703
G
T
R701
20k
R703
10k
U702.2
TL072
R702
1M
C703
100p
D703
C702
10n
R704
10k
R705
100k
+12V
C701
100n
U701.1
4001B
C704
100n
R706
100k
U702.1
TL072
D704
C705
1n
U701.2
4001B
U701.4
4001B
R1
A1
U701.3
4001B
D1
R707
20k
R708
1M
C707
100p
R709
10k
C706
100n
R710
20k
C708
100n
-12V
U703.1
40106B
U703.6
40106B
D705
C709
1n
R711
100k
R712
10k
U703.2
40106B
U703.3
40106B
U703.4
40106B
S704
Man Trig
R714
10k
R713
47k
C710
10n
U703.5
40106B
D706
R715
10k
D707
C711
100u 25V
C712
100n
U704.3
TL074
4066B
U705.4
R717
1k
U705.2
VR701
1M Log
Attack
Decay
VR702
1M Log
C713
100n
U705.1
U704.1
TL074
100k
VR703
Sustain
U704.4
TL074
R716
2k
C714
100n
R718
10k
U705.3
VR704
1M Log
Release
R720
1k
U704.2
TL074
R719
1k
ADSR1 Out
C715
1u Tant
C718
15u Tant
Time
S705
L
S
C716
100n
C717
100u 25V

Fig.63. *MIDI Ultimate* ADSR1 circuit.

ADSR circuit description

The ADSR circuit shown in Fig.63 is perhaps the most complex in this synth. (You can get simpler ADSR circuits that do not perform as well, or use dedicated ADSR chips). A philosophy for this synth was to use discrete components to ensure longevity through servicing with 'off the shelf' components. (Even that approach is now being tested by the decreasing availability of through-hole versions of many common chips that we have used for many years due to the move to surface-mount construction.)

Both ADSRs are identical in operation; the circuit in Fig.63 is for ADSR1. The ADSR has two modes of operation, namely gate and trigger. These terms are sometimes used interchangeably, but at Soundtronics, like many other synth designers we adopt the convention as best explained though the use of a keyboard:

- **Gate** – goes high when a key is pressed and remains high for as long as the key is pressed. The rising edge indicates that a key has been pressed, a falling edge that the key has been released
- **Trigger** – goes high when a key is pressed but almost immediately goes low again even though the key has not yet been released. The pulse width depends on the trigger source, but could be in the range of 50µs to 5ms, ideally less than 1ms.

Various manufacturers have taken different approaches to implementing gates and triggers, and their polarity. Some modular synth ADSRs respond simultaneously to both gate and trigger signals where the trigger can re-trigger the attack phase. For the *MIDI Ultimate*, we kept things simpler due to some constraints through the use of the toggle patching switches, but did include the mode option. The ADSR circuit is capable of re-triggering but not in this particular implementation. This is worth noting in case you decide to expand the *MIDI Ultimate* in the future with the addition of external modules.

Now that you have an understanding of the output envelope from the ADSR with its four distinct stages we shall move straight on with the circuit description.

Starting with the ADSR in Gate mode at S703 and with no gate pulse present, the ADSR is in its quiescent state with the output of U701.4 – point 'R1', high. Point 'R1' is high because both of its inputs are low. The high state of point 'R1' closes the bilateral switch U705.3, which discharges C715 to ground through the Release pot (VR704).

U702.1 is a comparator with the inverting input set at 2V by R706/R710 which determine the voltage threshold

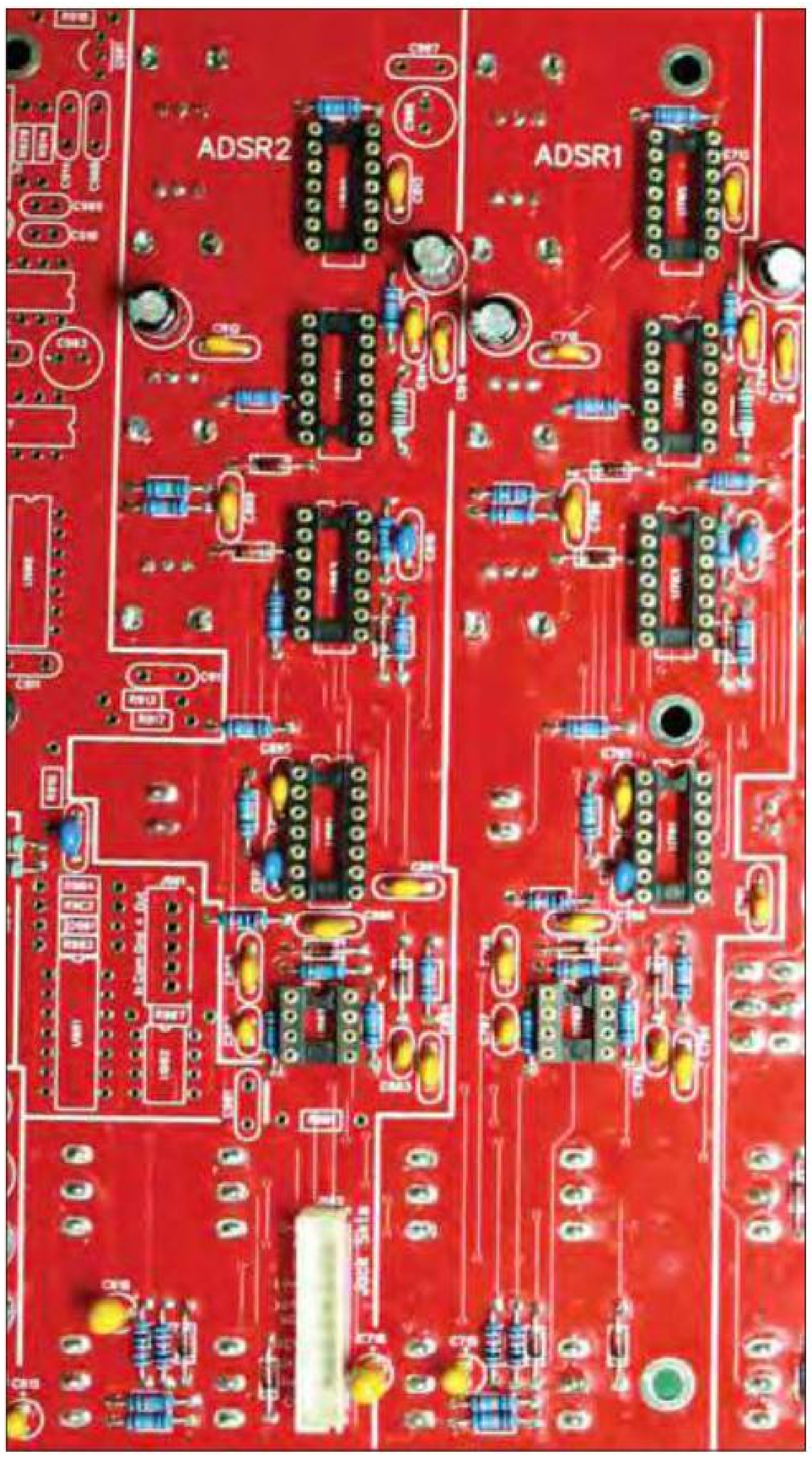


Fig.64. ADSR1 and ADSR2 assembled. (Whoops, I managed to solder the IC socket for U804 upside down!)

for the gate pulse at which the comparator switches state. When the gate pulse voltage exceeds 2V, the comparator U702.1 output will rapidly go from a negative to a positive voltage. This positive-going voltage via D704 creates a short duration pulse at pin 1 of U701.1. The pulse width is determined by the values of the RC circuit of C705 and R705. U701.1 and U701.2 form a flip-flop or latch circuit, with pin 1 being the set and pin 6 the reset. The pulse at pin one sets the flip-flop so its output at pin 4 of U701.2 – point 'A1' – goes high, which turns off bilateral switch U705.3 and turns on bilateral switch U705.2. C715 (and C718 if range switch is set to 'long') begins to charge up at a rate depending on the Attack pot (VR701). The voltage across C715 is buffered by U704.2 and connects to the ADSR1 output bus through R719. So, we are now in the attack phase with the period set by the RC circuit of VR701 + R720 and C715 (+C718).

The ADSR output voltage also feeds a comparator (U704.3) with its threshold set to 10V by U704.4, R716 and R718. When the attack voltage exceeds 10V, the output of U704.3, pin 8, quickly goes from a negative state to a positive state. This positive-going signal resets (pin 6) the flip-flop of U701.1 and U701.2 with its output (pin 4) point 'A1' going low. As the gate signal is still present at pin 13 of U701.4 via Schmitt inverters U703.1 and U703.6, point 'R1' remains low, but with point 'A1' going low, bilateral switch U705.2 opens. These conditions also cause U701.3 output, pin 10 (point 'D1') to go high indicating that we are now in the decay phase.

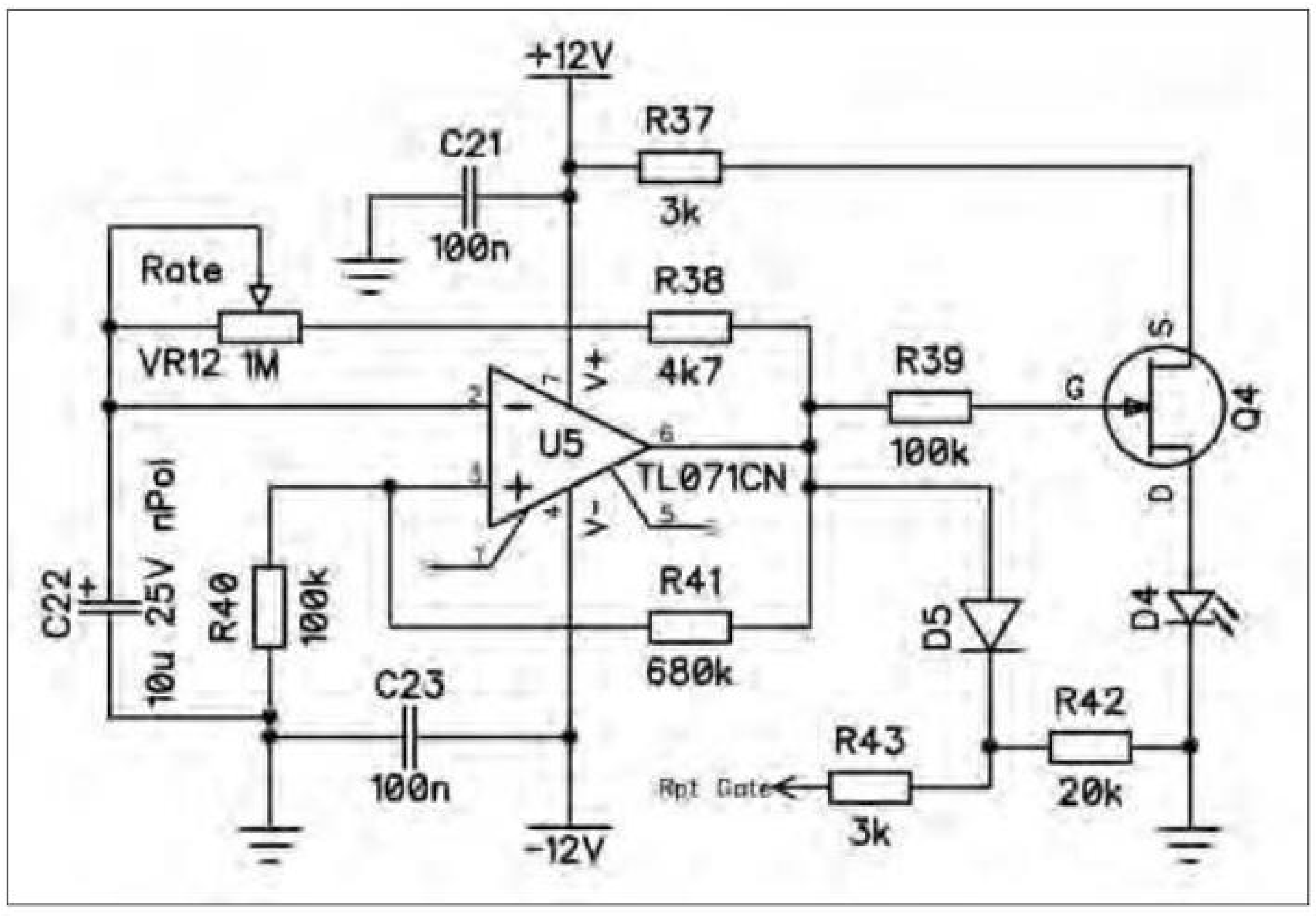


Fig.65. *MIDI Ultimate* repeat gate circuit.

With point 'D1' high, bilateral switch U705.1 closes, which discharges C715 (plus C718) through the decay pot (VR702 plus R720). This time, C715 (plus C718) will only discharge to the voltage level at the output of U704.1, which is the sustain level set by VR703. This is now a steady-state condition and will remain so until the gate pulse is removed, at which point C715 (plus C718) is discharged to ground when bilateral switch U705.3 closes through the release pot (VR704 plus R720).

If the gate is applied and release before the decay phase, the output of U703.1 goes high and resets the flip-flop, so that the release phase is immediately commenced.

If the mode is changed to trigger at S703, a trigger pulse is generated with U702.2 and the flip-flop latches as in the gate mode. However, on completion of the attack phase and as there is no gate signal present, the release phase is executed. Therefore, in trigger mode the ADSR is actually an AR envelope generator with the period independent of how long a key is present.

Repeat gate circuit description

The repeat gate generator in Fig.65 is a square wave oscillator used to continually trigger the ADSR envelope generators, especially in trigger mode.

U5 has positive feedback applied through R41 to its non-inverting input. On power up, U5's output may be high or low as it is indeterminate, but let's assume low at around –11V. C22, which is a non-polarised aluminium capacitor and was at ground potential begins to charge towards –11V through VR12 plus R38. A potential divider of R41 and R40 hold the non-inverting input at around –1.4V. Eventually, C22 charges up a voltage lower than –1.4V and as U5 acts as a comparator, its output flips high to +11V. Now the voltage at the non-inverting input is held at +1.4V and C22 begins to charge towards +11V. Once the voltage across C22 exceeds +1.4V, U5 flips low again and the cycle repeats – oscillates. LED D4 driven by Q4 indicates the state of the pulse output.

Sample and Hold (S&H) circuit description

The S&H circuit of Fig.67 consists of two sections, the sample and hold plus a pulse generator. The pulse generator is very similar to the repeat gate generator described above. The main difference is the introduction of D1, which quickly charges C17 through R24 when U3 output is high, bypassing the sample rate pot (VR9). Unlike the square wave output from the repeat gate generator, the S&H clock is a pulse waveform with a short positive-going pulse with a frequency range of around 0.5 to 50Hz.

The S&H section has a choice of two input signals to sample, LFO2 output and the white noise source, selected by S1. The selected input signal is applied to the input level pot (VR11) and from there to the inverting input of op amp U4.1 via R27. U4.1 inverts this signal with a gain of 0.5 and offsets the voltage to a level of about 6.5 to 7V. The output of U4.1 is applied to the sampling FET transistor switch (Q2). Q2 provides a higher off resistance when the signal on the source pin 'S' is kept above ground potential.

The pulse output from the clock generator is applied via the differentiator circuit of C11 and bias network of R22 and R23 to the gate pin of Q2 to shorten

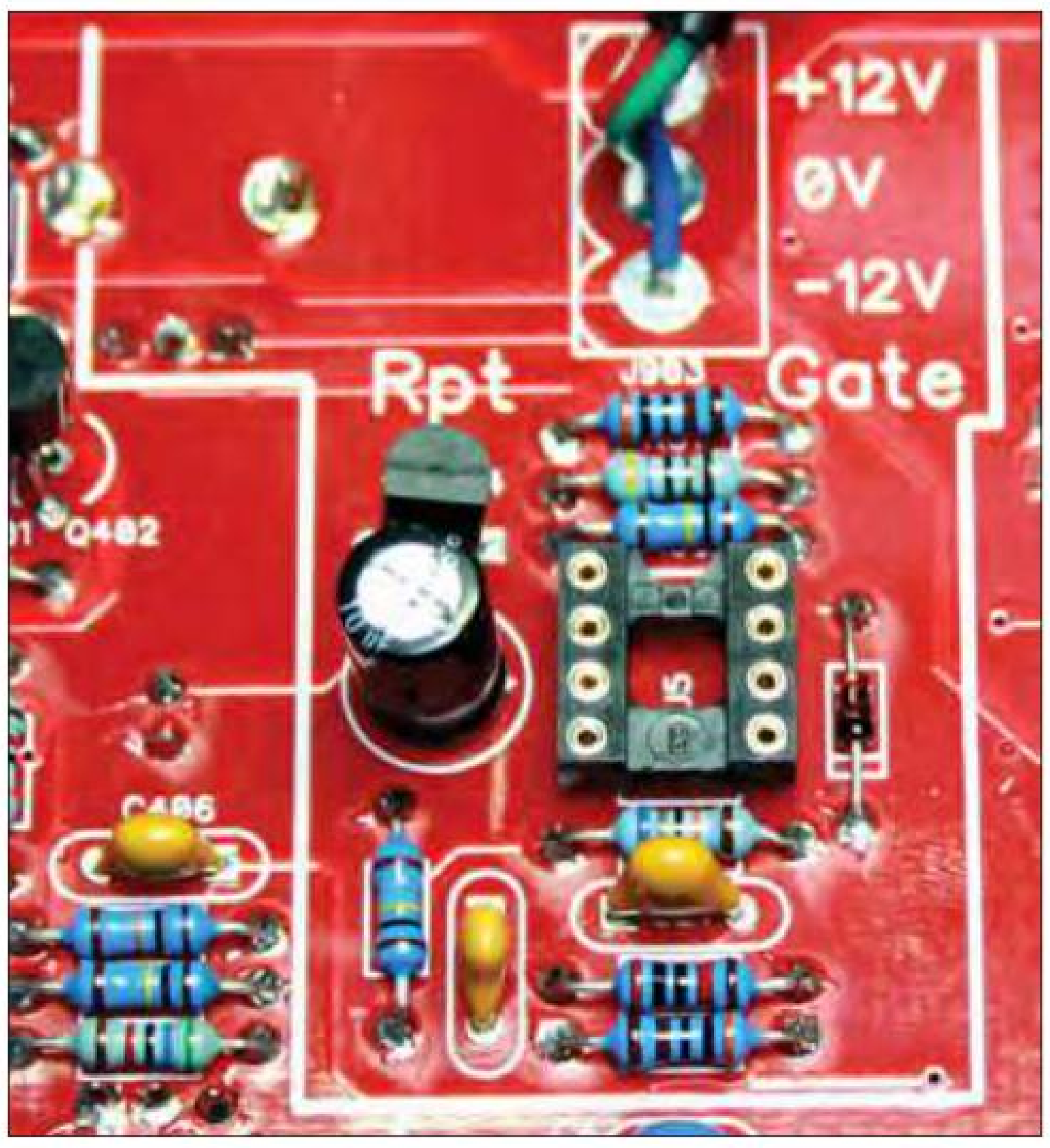


Fig.66. Repeat gate assembled.

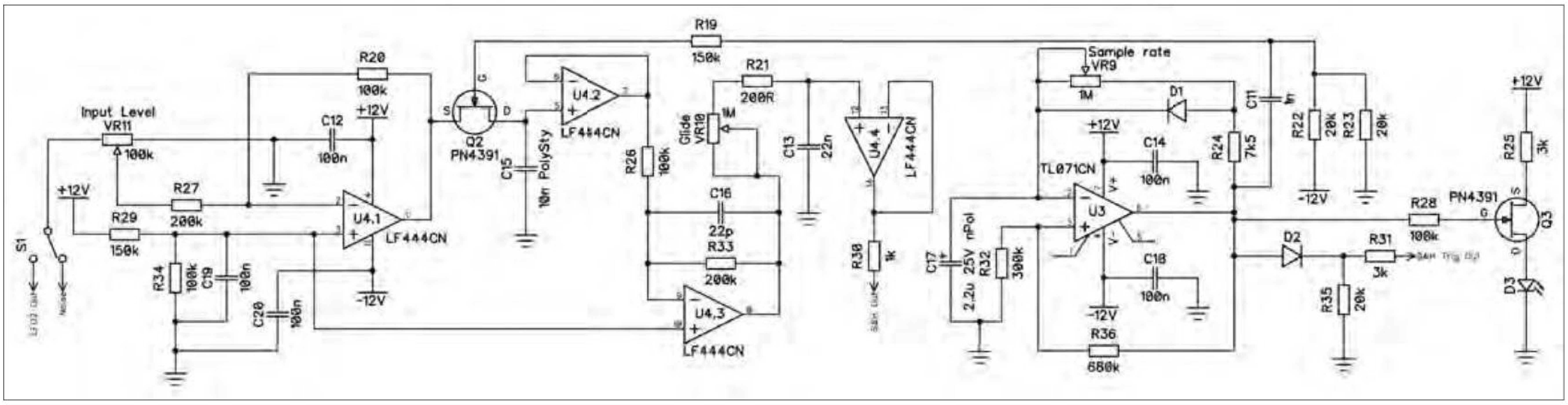

R19
150k
R20
100k
R21
200R
Input Level
VR11
100k
C12
100n
+12V
+12V
R27
200k
R29
150k
S1
LFO2 Out
Noise
U4.1
LF444CN
R34
100k
C19
100n
C20
100n
-12V
S
G
D
Q2
PN4391
C15
10n PolySty
U4.2
LF444CN
R26
100k
Glide
VR10
1M
C16
22p
R33
200k
U4.3
LF444CN
C13
22n
U4.4
LF444CN
R30
1k
S&H Out
Sample rate
VR9
1M
D1
C11
1n
+12V
TL071CN
U3
V+
V-
C14
100n
C18
100n
-12V
C17
2.2u 25V nPol
R32
300k
R36
680k
R24
7k5
R22
20k
R23
20k
-12V
R25
3k
+12V
R28
100k
PN4391
Q3
D2
R31
3k
S&H Trig Out
R35
20k
D3

Fig.67. *MIDI Ultimate* S&H (sample and hold) circuit.

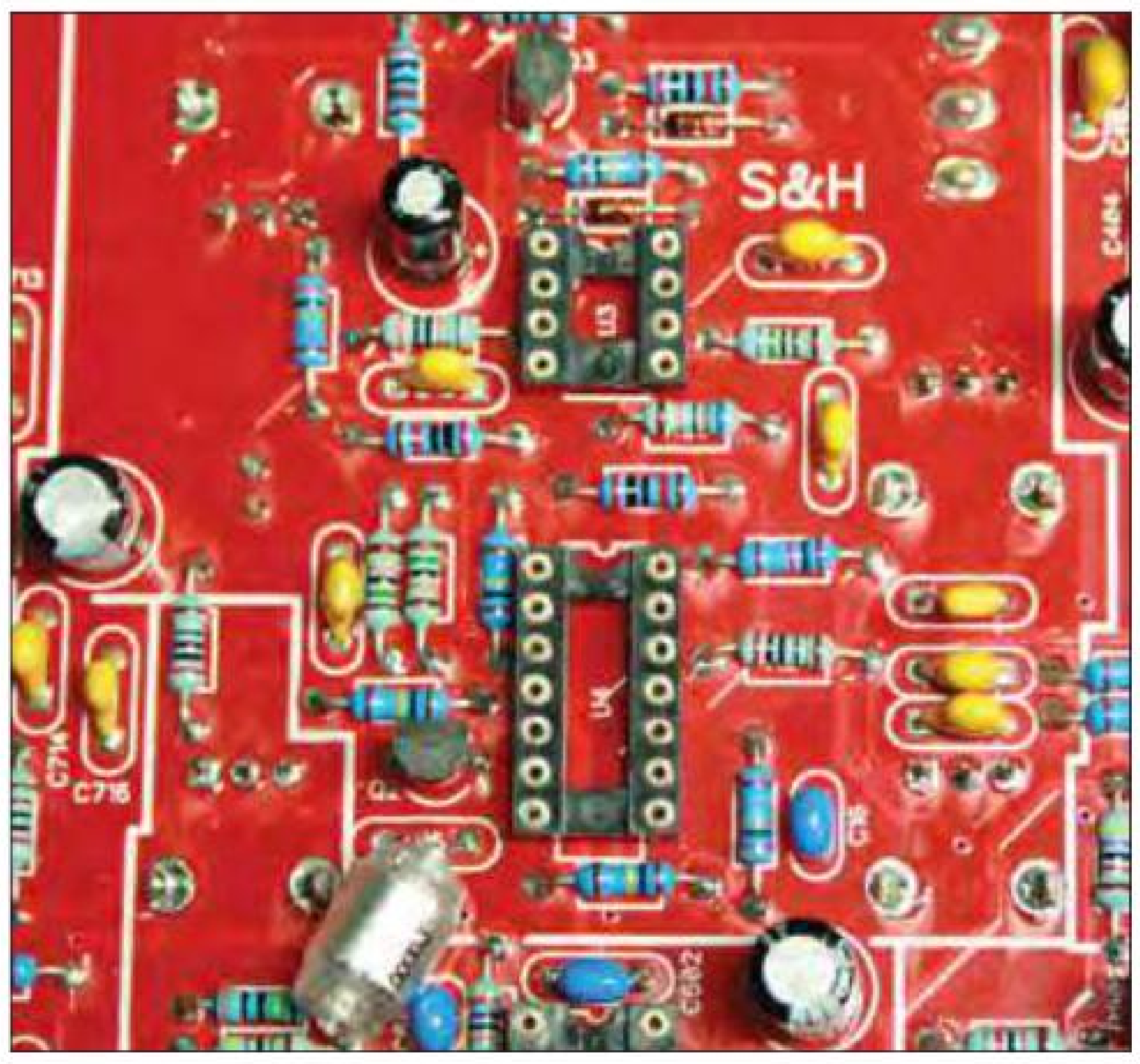


Fig.68. S&H (sample and hold) ssembled.

the pulse width even further. This ensures Q2 is only switched on for the briefest of time and effectively connects the output of U4.1 to C15 which charges up to the same voltage level as U4.1 output. C15 is a low-leakage polystyrene capacitor, chosen because we need it to retain or 'remember' that voltage level for up to two seconds. After the gate pulse, the gate voltage returns to around –6V bias from R22 and R23, which effectively makes Q2, open circuit between the source and drain pins. This means C15 has nowhere to discharge to other than the non-inverting input of U4.2. However, U4.2 has a high input impedance of 100s of megohms, thus preventing any significant voltage droop. The sampled voltage from U4.2 is re-inverted again by U4.3, which also has a gain of 2 to restore the voltage to the original level. The output of U4.3 is fed to an RC circuit consisting of glide pot VR10, R21 and C13 which is buffered by non-inverting op amp U4.4 The output of U4.4 feeds the S&H bus via R30. D3 indicates the clock sample rate and the clock is also available on the S&H gate bus, which can also be used to trigger the ADSRs.

You should now have completed the assembly of the modulation sources, and the pot side of your PCB should look like Fig.69.

We have had a request for a PCB layout drawing showing the component identification because the number is hidden once the component is soldered in. A PDF is available from the Soundtronics website and from the *Practical Electronics* website.

Testing the modulation sources

Inserting chips for each new section one at a time for testing before moving onto the next is recommended but not essential.

As the patch switches allow for numerous routing options, the focus here is to prove the functionality of the modulation sources assembled this month using VCO1. Once proven, you should then try out similar tests with the other two VCOs, VCF and VCA.

Before you start testing, set all 3-way switches to centre off, all 2-way switches to the down position, all pots fully counter-clockwise except the master level, which you should adjust to a comfortable level. You will also need the headphones connected as described in *Part 3*, Fig.58. In table 4, some pot settings are referenced to clock setting because at this point, you have no front panel scales to use. Each pot has a slot in the end that you can use as a pointer.

None of the modulation sources covered in this part requires calibration. The VCOs and VCF calibration are best left until the MIDI interface is

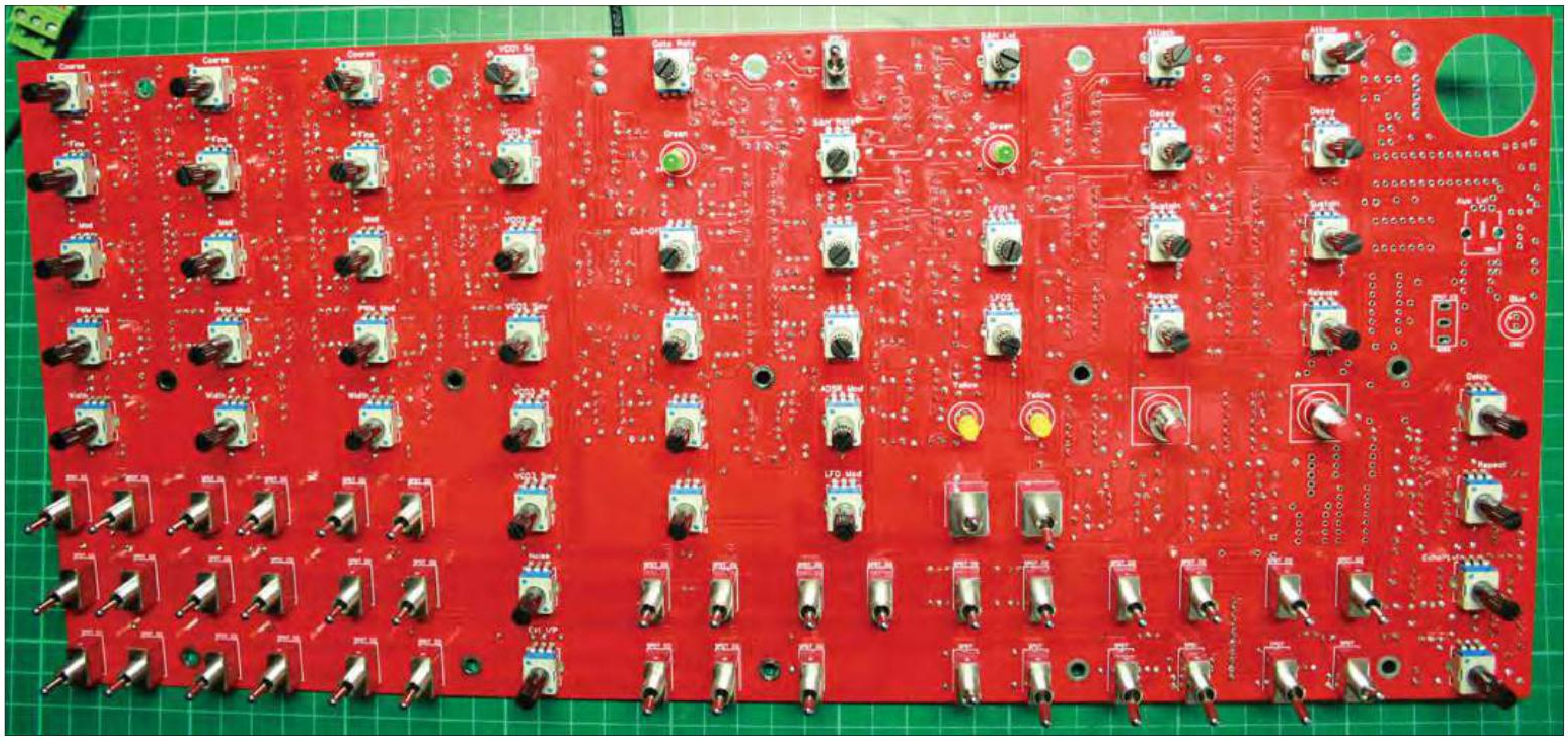


Fig.69. PCB assembled and ready for testing.

Table 4: Schedule of tests to prove functionality of the modulation sources – LFOs, ADSRs, Repeat Gate and S&H.

No.	Test	Pots	Switches	Result
1	Establish a base point	VCO1 coarse frequency 12 o'clock, mixer for VCO1 square 100%, VCF filter cut-off 100%, VCA initial level 100%		You should hear a tone from VCO1
2	Prove LFO1	VCO1 modulation 9 o'clock, LFO1 rate 9 o'clock	VCO1 – LFO1 Mod, LFO1 – Square	'Nee-nah' tone
Vary LFO1 rate for the speed of the nee-nah. Switch LFO1 square to ramps and try the different ramp waveforms				
3	Prove LFO2	VCO1 modulation 12 o'clock, LFO2 rate 12 o'clock	VCO1 – LFO2 Mod, LFO2 – Square	Nee-nah tone at a faster rate than test 2
As for LFO1, carry out same tests for LFO2 then return switches and pots to their position as in test 1				
4	Prove ADSR2	VCO1 modulation 9 o'clock, ADSR2 attack 3 o'clock, decay 12 o'clock, sustain 12 o'clock, release 3 o'clock	VCO1 – ADSR2, Press, hold then release ADSR 2 manual gate button	Tone will increase in frequency, then decrease and remain steady until gate button released, then the frequency will decrease to the starting point
Return switches and pots to their position before test 1				
5	Prove ADSR1 and Repeat Gate	VCO1 coarse frequency 12 o'clock, mixer for VCO1 square 100%, VCF filter cut-off 9 o'clock, VCA initial level 100%. ADSR1 attack, decay, and sustain to 12 o'clock, release 3 o'clock. VCF ADSR mod fully clockwise	ADSR1 Repeat Gate, VCF ADSR1	Repeating dull to bright sound
6	Prove S&H gate	S&H sample rate to 12 o'clock	ADSR1 gate to Trig mode, ADSR1 S&H Gate	Pulsating sound
7	Prove S&H	S&H input level to 3 o'clock.	VCO1 to S&H Mod	Random notes

available, but we can now set up the VCA as we have the LFOs available.

VCA Calibration

In *Part 3*, we constructed the VCA (Fig.53), now we can calibrate it using the two presets P501 and P502 and LFO1.

1. Turn all mixer controls to minimum, LFO1 set to square wave, its frequency set to maximum with the range switch set to 'high'. VCA modulation pot set to maximum. VCA modulation switches, LFO1 selected, other two switches centre off.
2. Attach your scope (see *Part 2*, page 41 on how to make your own PC soundcard oscilloscope) to the output of U501.1 pin 8 where it links to the top of R501, or plug your headphones in with the volume level set to maximum. There are numerous 0V ground points but I find clipping a lead to one of the mounting holes works just fine.

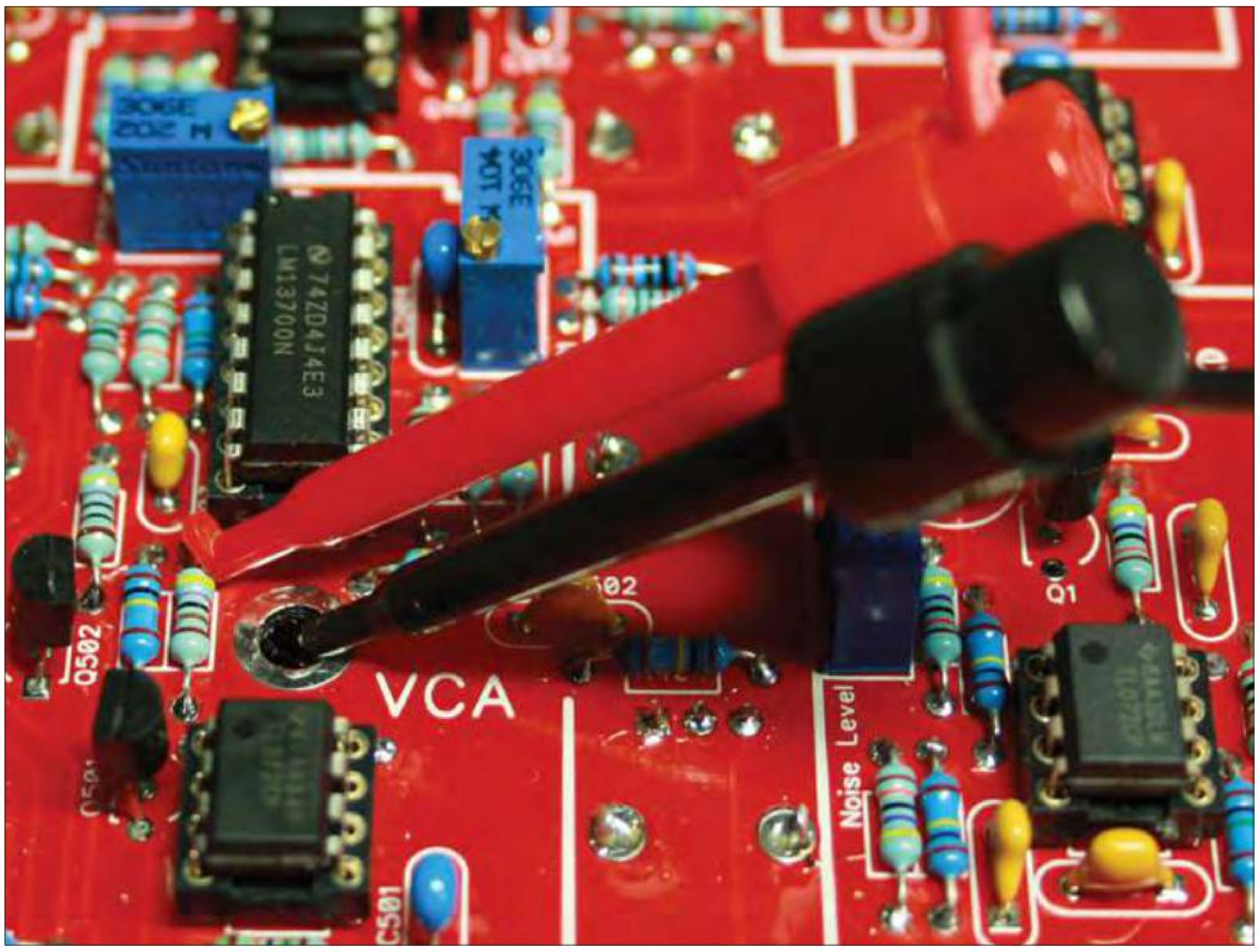


Fig.70. Test points for setting up P501 and P502 presets.

3. The adjustment we are making is to ensure the minimum breakthrough of the control voltage onto the audio output. Adjust the CV Null preset P502 until the amplitude of the LFO1 square wave is at a minimum, which should be less than 5mV. If using headphones, adjust P502 until the audible tone from LFO1 is at its quietest.
4. Switch the VCA modulation switch from LFO1 to centre off. Connect a DC voltmeter to the VCA output at the same point as the scope was connected to and adjust P501 until the output voltage is as close to 0V as you can make it. P501 can easily adjust to around 0.1mV, but that depends on the resolution of your voltmeter.

Last, but not least this month, here is some inspiration for constructors – a link to a patch which I have called 'Pulsing filter': **https://youtu.be/kjMuVyiLIP8** – worth playing back through some big speakers.

Next month

You now only have one section remaining to build on the main PCB, which is the MIDI interface – so you can now really start to find your way around the synth.

Next month, is all about MIDI, both as a general topic and how it is implemented within the *MIDI Ultimate.*

Practical Electronics

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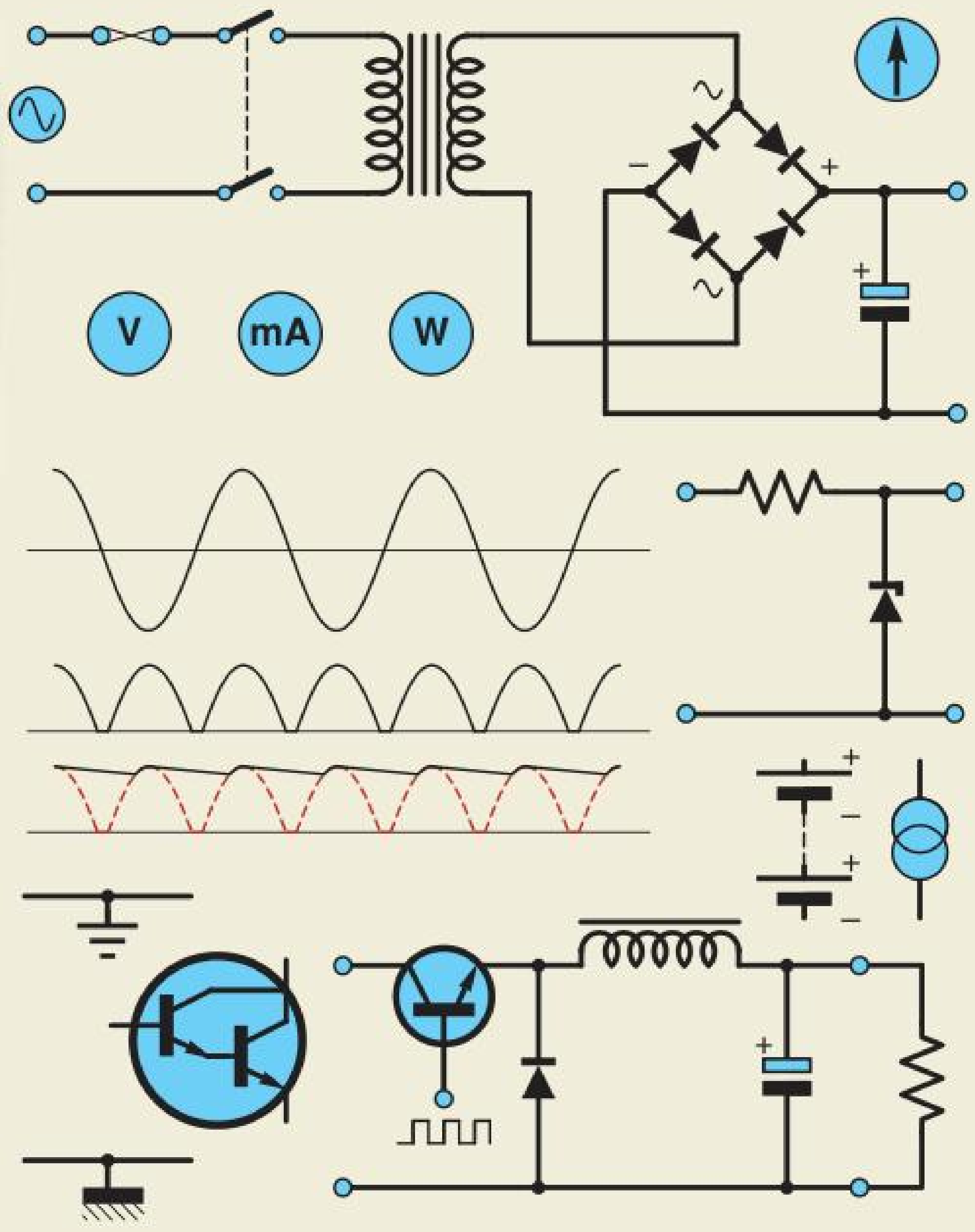
Teach-In 2019

Powering Electronics

Part 6: Voltage multipliers



by Mike Tooley

Your project is finished and ready to go, but the job isn't done until you've found an appropriate source of power. This could be as simple as choosing a suitably rated mains adapter or as complex as designing a switched-mode power supply with multiple outputs and battery backup. Our latest series – *Teach-In 2019* – is here to help, and will provide you with insight into all aspects of powering your electronic projects and designs.

Last month, we examined switched-mode power supplies (SMPS) and explained the basic principles of buck (step-down) and boost (step-up) switching regulators. This month, we will be revisiting some of the rectifier circuits that we first met in *Part 2*, and show how we can easily produce multiples of the basic DC output voltage with voltage doublers, triplers, and quadruplers. These are useful circuits that can be built using just a handful of low-cost components.

We will also provide some important advice and guidance on how to select and use electrolytic capacitors.

Our *Practical Project* takes the form of a simple voltage doubler designed for producing a low-current 15V DC output from a small 9V battery. This project is ideal for supplying power to 15V CMOS logic circuits.

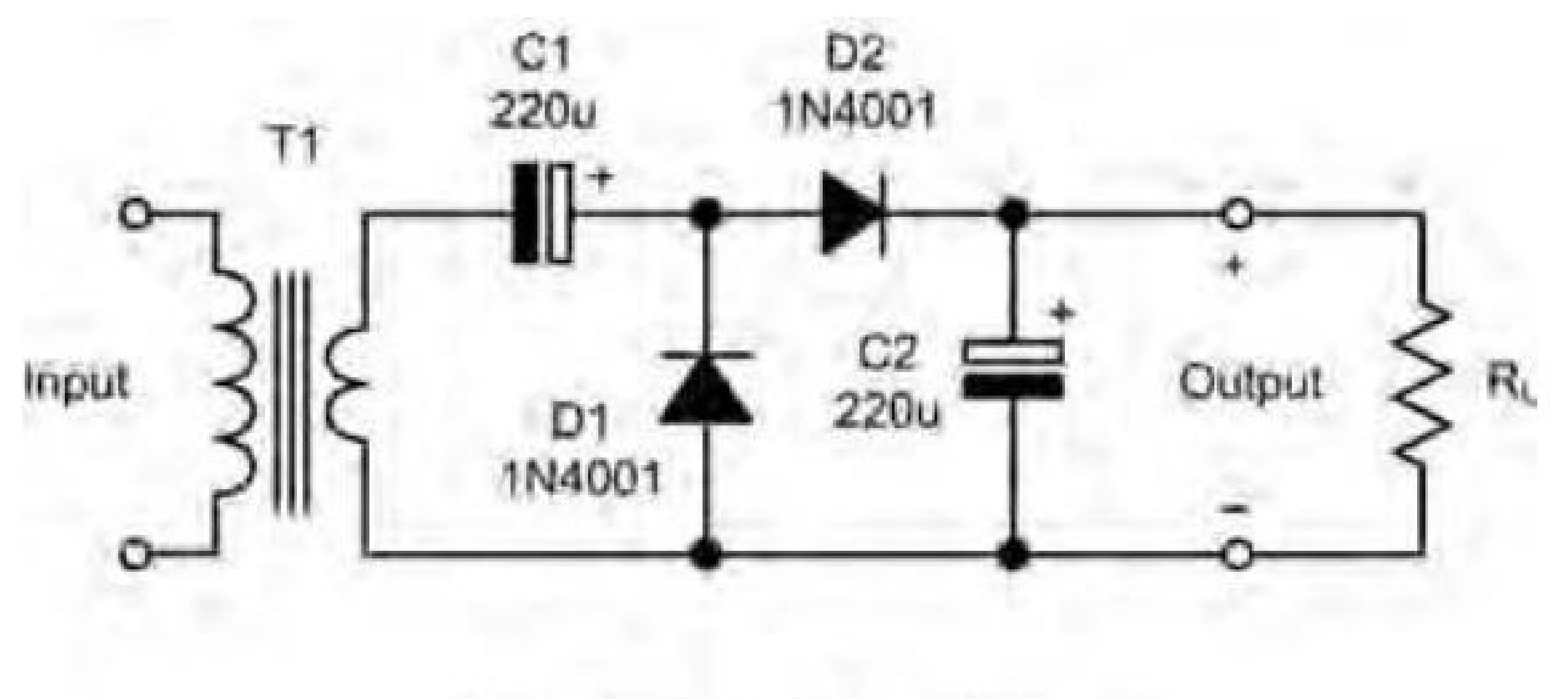
Voltage multipliers

The simple half-wave and full-wave rectifiers that we met in *Part 2* can be easily modified and extended to produce multiples of the basic DC output voltage. For example, Fig.6.1(a) shows a simple half-wave voltage doubler arrangement. On the first negative-going half-cycle of the input AC supply voltage, D1 will conduct and charge C1 to the negative peak secondary voltage (less the forward 0.7V voltage drop associated with D1). This positive voltage appears at the anode of D2, which then conducts on the positive-going half-cycle, charging C2 to *double* the applied peak voltage (less the forward voltage drop associated with D1 *and* D2). With a sinusoidal AC input, two 1N4001 diodes, and an RMS secondary voltage of V_S, the peak DC voltage appearing across the load (R_L) will be approximately:

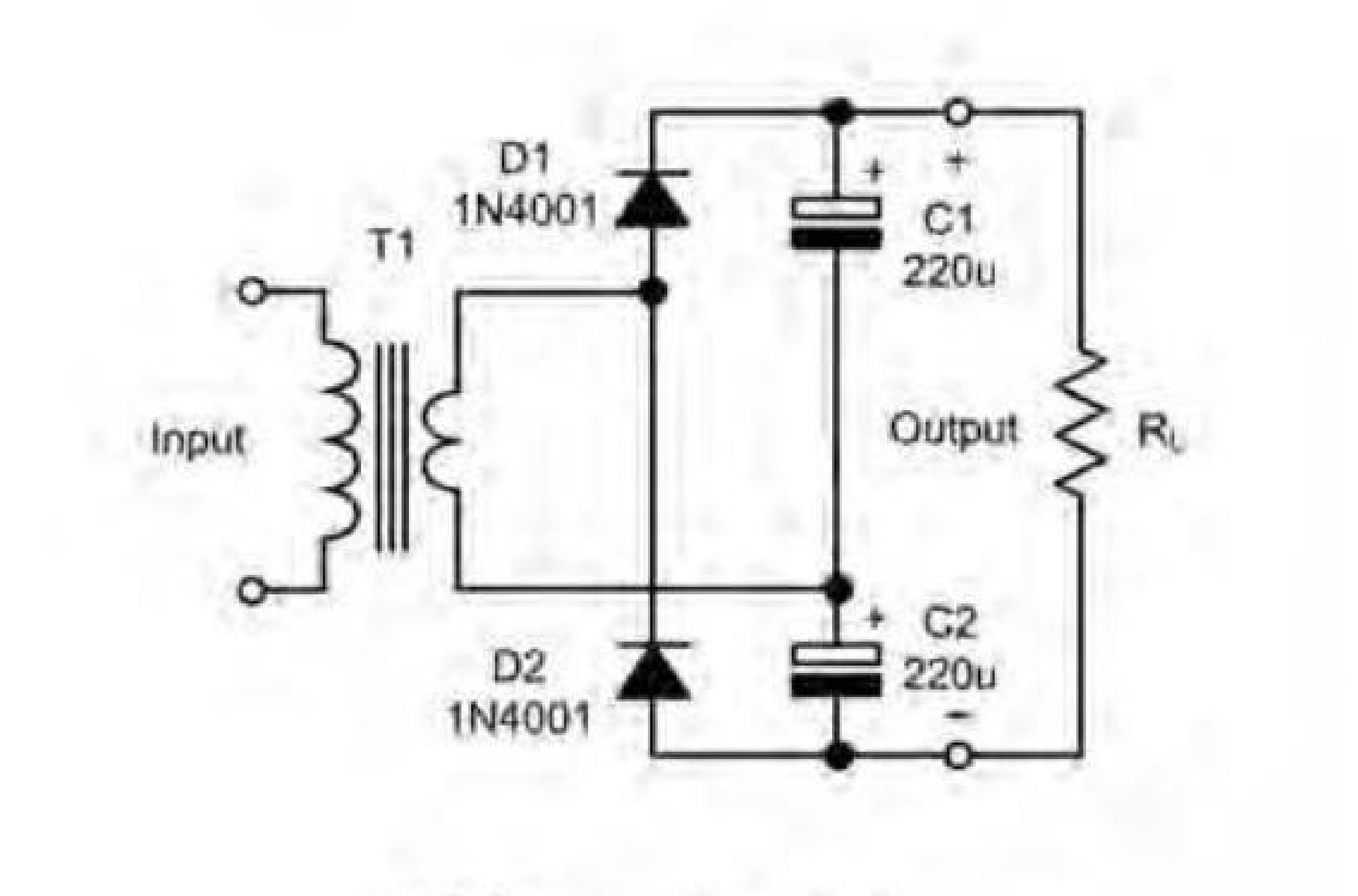
$$V_{out} = (2 \times 1.4 \times V_S) - (2 \times 0.7V) = 2.8V_S - 1.4V$$

Note that the 1.4 factor converts the RMS value to a peak value (strictly speaking the factor is √2 = 1.4142…). So, with a transformer delivering 12V RMS from its secondary, the no-load output voltage can be expected to be approximately 32V (and also at low load currents of up to about 50mA).

An alternative voltage doubler arrangement is shown in Fig.6.1(b). This circuit offers better performance to that shown in Fig.6.1(a) since both diodes are conducting for part of every half-cycle of the secondary input (its action is that of a

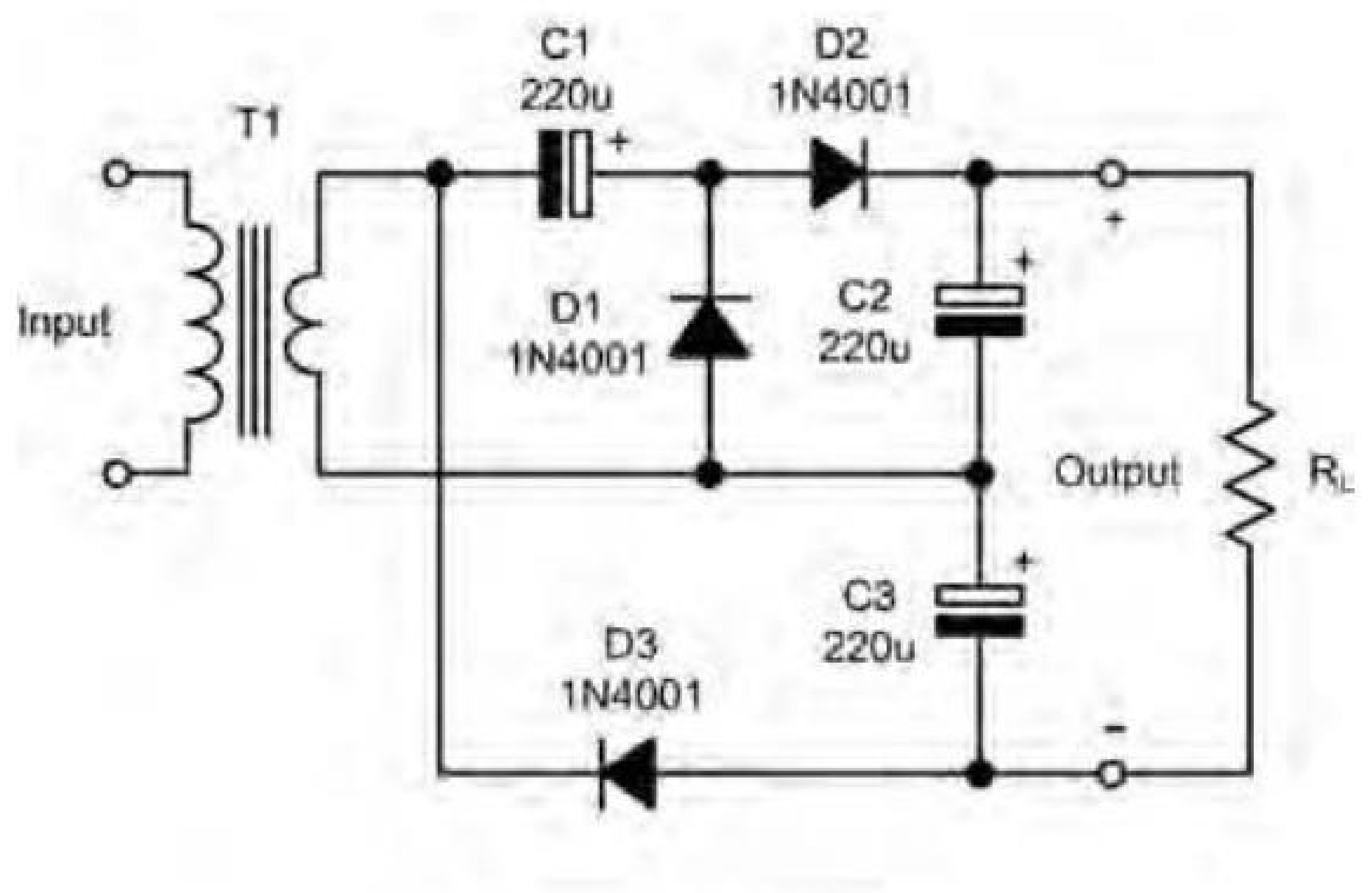


(a) Half-wave voltage doubler

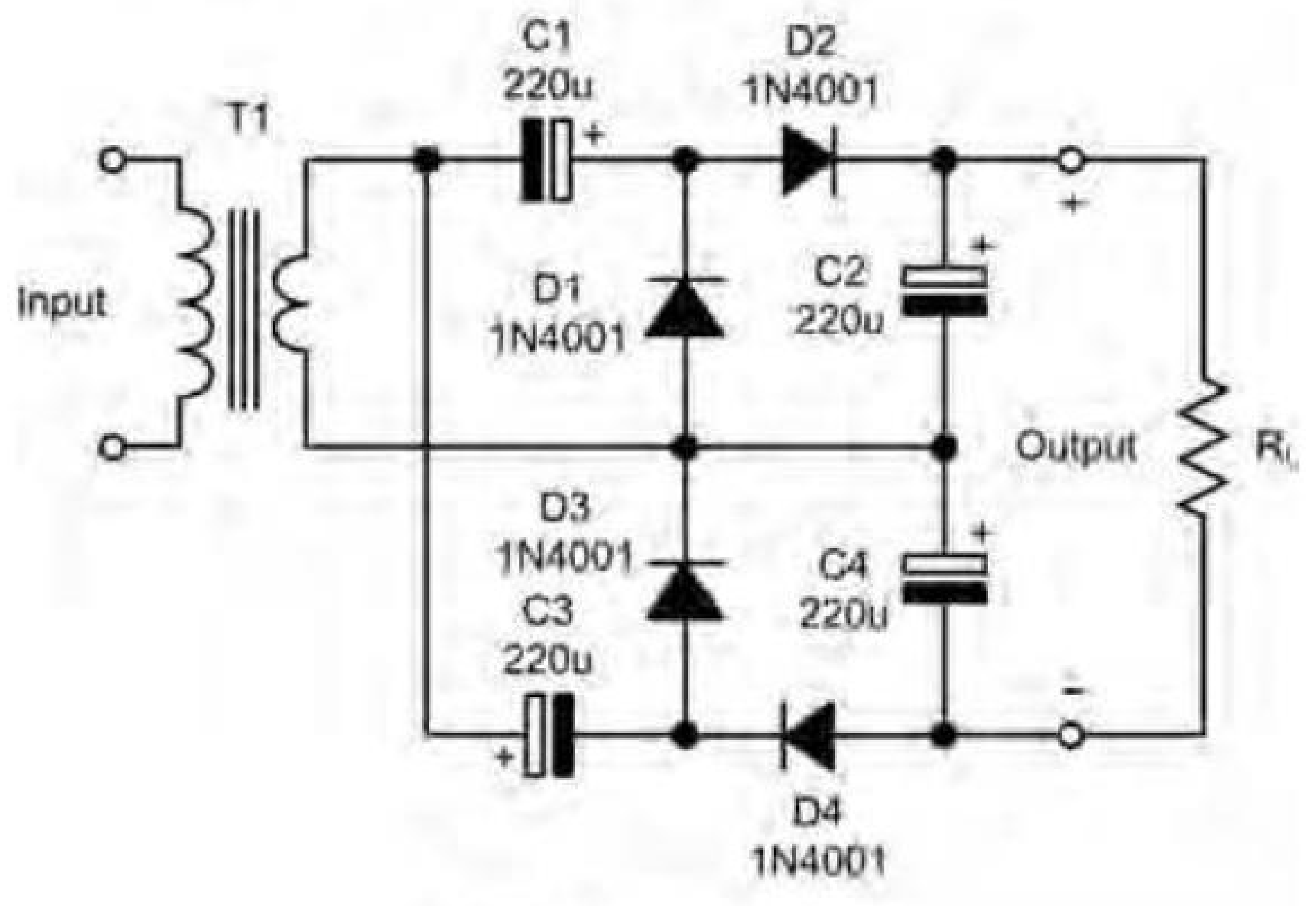


(b) Full-wave voltage doubler

Fig.6.1. Simple half-wave and full-wave voltage doublers.



(a) Voltage tripler



(b) Voltage quadrupler

Fig.6.2. Voltage tripler and voltage quadrupler arrangements.

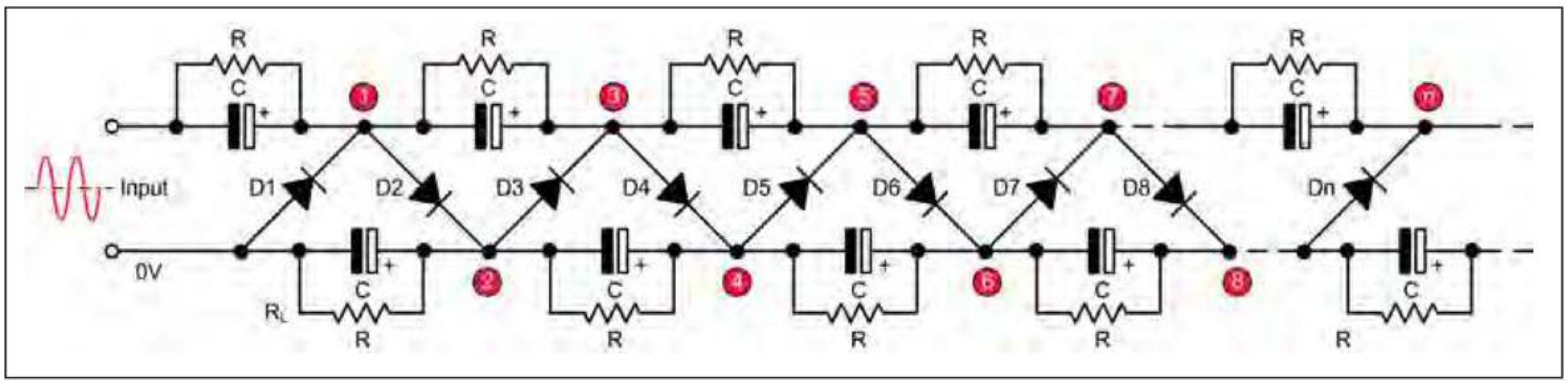
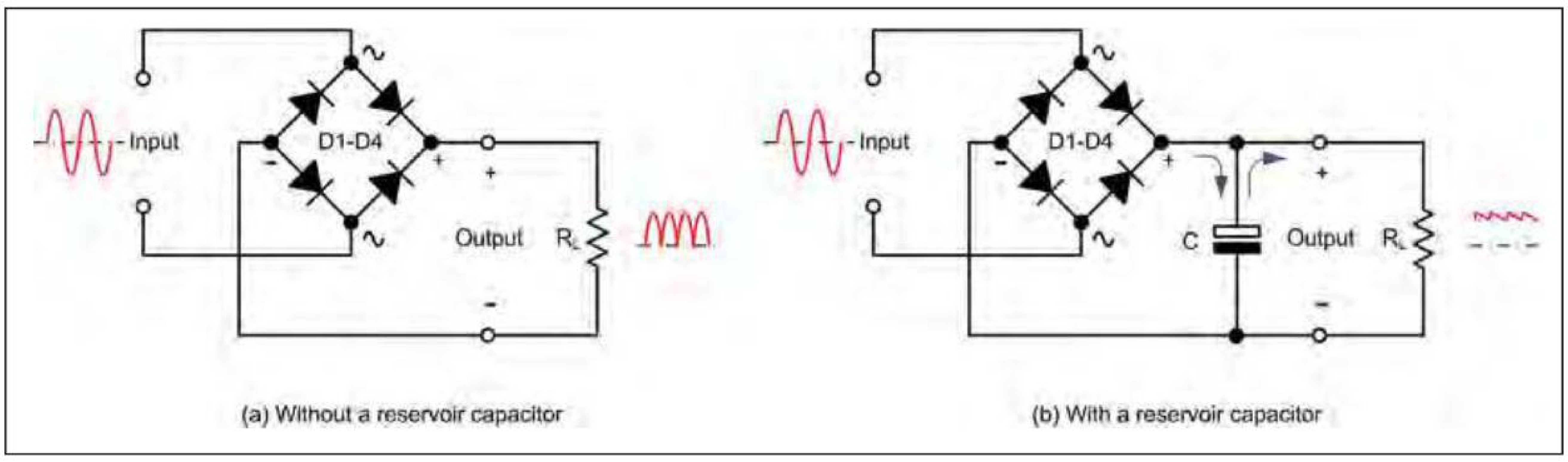


Fig.6.3. Voltage multiplier Cockcroft-Walton 'n-tupler' ladder.



(a) Without a reservoir capacitor

(b) With a reservoir capacitor

Fig.6.4. A bridge rectifier with, and without, reservoir capacitor. Note how the current flows alternately into and out of the capacitor in order to maintain current through the load.

full-wave rectifier rather than a half-wave rectifier). With a 24V RMS secondary, this circuit will deliver approximately 65V at load currents of up to 50mA.

Fig.6.2(a) shows a simple voltage tripler. The upper part of this circuit (C1/C2 and D1/D2) forms a voltage doubler with double the peak secondary voltage appearing across C2. To this is added the negative peak voltage produced by the rectifying action of D3 which charges C3 on negative-going half cycles. Thus, the voltage appearing across the load (R_L) will be approximately three times the peak secondary voltage delivered by T1.

With a sinusoidal AC input, three 1N4001 diodes, and an RMS secondary voltage (V_S) the peak DC voltage appearing across the load (R_L) is approximately given by:

$$V_{out} = (3 \times 1.4 \times V_S) - (3 \times 0.7V)$$
$$= 4.2V_S - 2.1V$$

Thus, with a transformer delivering 24V RMS from its secondary, the no-load output voltage will be approximately 98V and also with load currents of up to 50mA.

Fig.6.2(b) shows a simple voltage quadrupler. The upper part of this circuit (C1/C2 and D1/D2) forms a voltage doubler with double the peak secondary voltage appearing across C2.

The lower part of the circuit (C3/C4 and D3/D4) forms a further *negative* output voltage doubler, with its output appearing across C4. The resulting output voltage delivered to the load (R_L) is the sum of the voltages appearing across C2 and C4. With a sinusoidal AC input, four 1N4001 diodes, and an RMS secondary voltage (V_S) the peak DC voltage appearing across the load (R_L) will be approximately:

$$V_{out} = (2 \times (2 \times 1.4 \times V_S)) - (4 \times 0.7V)$$
$$= 4V_S - 2.8V$$

With a transformer delivering 24V RMS from its secondary, the no-load output voltage will be approximately 130V, which can be maintained at load currents of up to 50mA.

Higher voltages

It is possible to generate *much* higher voltage using multiplier arrangements, but at the cost of poor regulation and limited output current. The general arrangement of an *n*-tupler ladder voltage multiplier is shown in Fig.6.3. Originally devised by John Cockcroft and Ernest Walton more than 90 years ago, this circuit was employed in a high-voltage DC power supply for use with a nuclear particle accelerator.

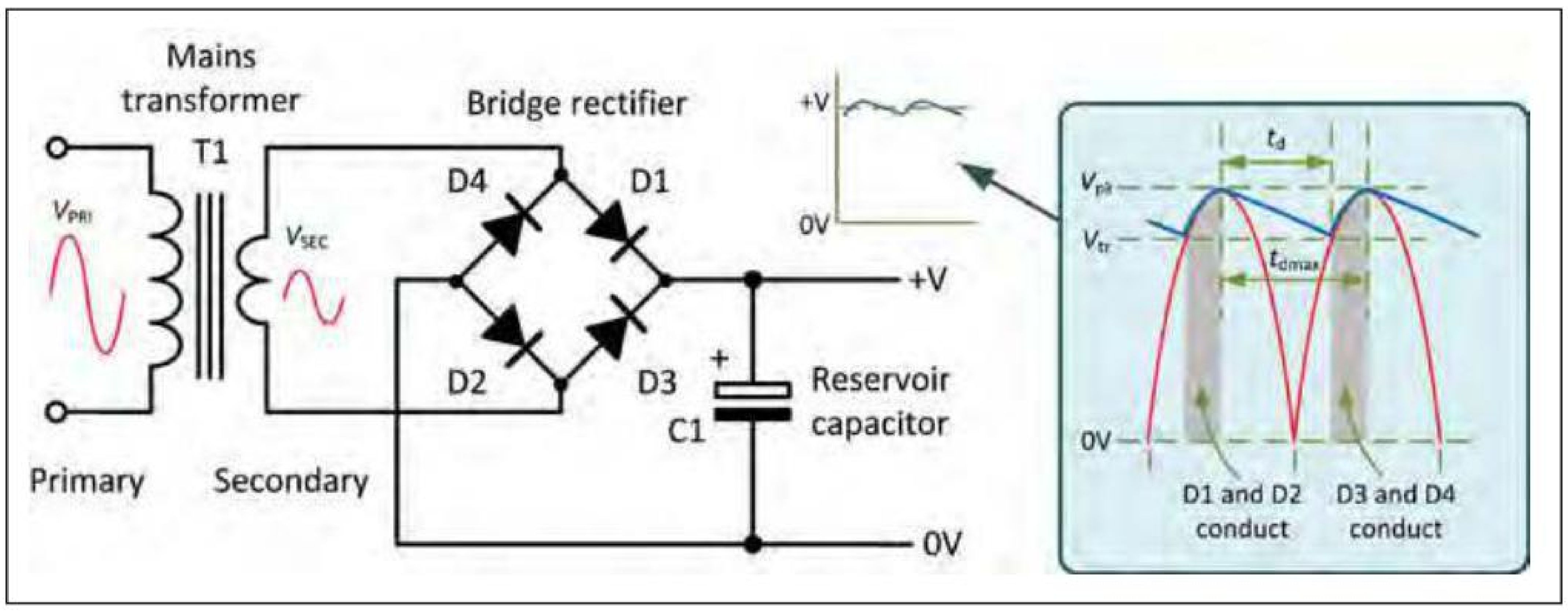


Fig.6.5. Effect of a reservoir capacitor on the output waveform of a full-wave bridge rectified DC power supply.

The Cockcroft-Walton arrangement cascades ladder networks of rectifiers and reservoir capacitors, and offers the advantage of not requiring a bulky and expensive high-voltage transformer. Note that in Fig.6.3 we have added a shunt resistor in parallel with each reservoir capacitor. These not only assist with equalising the voltages developed across the reservoir capacitors, but also provide a discharge path for the capacitors, making the circuit safe to handle when the power is no longer applied. Typical values for the shunt resistors are in the range 47kΩ to 1MΩ.

It should perhaps be emphasised that a high-order Cockcroft-Walton power supply is only capable of delivering relatively small currents. For example, a nine-times multiplier ($n = 9$) will only be capable of delivering a few tens of milliamps. This may be sufficient for a few applications but wholly insufficient for many others.

Reservoir capacitors

In both switched-mode and conventional linear supplies, the reservoir capacitors have the important role of 'holding-up' the output voltage from the bridge rectifier. The capacitors will charge on each peak of the applied voltage (during which the respective rectifier diodes will be conducting) and it will retain this charge when the secondary voltage falls (see Fig.6.4). The general rule of thumb is to use a capacitor, suitably rated in terms of working voltage and ripple current, with a value that is as large as possible within the physical constraints of the board on which it is mounted.

It is possible to obtain a rough 'ball park' value of the capacitance required by assuming that the capacitor has a maximum discharge time equivalent to the time for one half-cycle of the AC supply (recall that the bridge rectifier is providing full-wave rectification, and so the 'ripple' frequency will be 100Hz, *twice* the 50Hz frequency for a standard UK mains supply). The recommended value of capacitor can then be estimated from the relationship:

$$C = \frac{I_{max} \times t_{dmax}}{V_{pk} - V_{tr}}$$

Where I_{max} is the maximum load current, t_{dmax} is the maximum discharge time, V_{pk} and V_{tr} are the peak and trough voltages respectively (see Fig.6.5) and the DC output voltage (V_{DC}) is simply the average of these two.

Let's take a simple example based on a reservoir capacitor that what we might need for a basic audio amplifier operating from a full-wave bridge rectified 12V RMS AC supply and requiring a maximum current of 1A. The 12V RMS supply derived from a mains transformer secondary will have a peak value of approximately 17V (12V × 1.4). Let's assume that we would not be prepared to tolerate a voltage fall of more than 1V on the DC supply rail when the maximum load current is being supplied. We should thus plan for a worst-case situation in which $V_{pk} = 17V$ and $V_{tr} = 16V$ (in other words, we are designing for a maximum *ripple* voltage of 1V peak).

With a 50Hz supply the value of t_{dmax} would be 10ms and the maximum load current will be around 1A. Hence:

$$C = \frac{1\times10\times10^{-3}}{17-16}\,\mathrm{F} = 10\mathrm{mF} \text{ or } 10,000\mu\mathrm{F}$$

As a general rule, reservoir capacitors should have a maximum working voltage rating of at least 50% more than the expected DC voltage and with a ripple current that is at least 50% more than the maximum DC load current. Thus, a component rated at 30V or 35V with a ripple current rating of 1.5A, or more, would be suitable.

In this simple example we considered the bridge rectifier to be perfect and didn't account for any voltage drop that might be attributable to the rectifier diodes when in the conducting state. In some cases (and for low-voltage supplies in particular) it would be wise to consider the *additional* voltage drop that is developed across each semiconductor junction when it is conducting. We will illustrate this with a slightly more complex example.

Let's assume that we are using a linear three-terminal 5V regulator based on a 7805 and a mains transformer with an 8V secondary and a bridge rectifier arrangement (as before). The peak voltage appearing across the reservoir capacitor will be (8 × 1.41) = 11.3V *less* the forward voltage drops associated with *two* of the four rectifier diodes (only two of the four diodes will be conducting at any time). Thus, the peak voltage appearing across the reservoir capacitor will not be 11.3V but nearer to (11.3 – (2 × 0.7)) = 9.9V.

The minimum input voltage for the 7805 regulator is specified as 7.5V (see Table 3.1 on page 45 of *EPE*, February 2019). Thus, for a maximum load of 1A, we can now estimate the minimum required value of reservoir capacitance from:

$$C = \frac{1\times10\times10^{-3}}{9.9-7.5}\,\mathrm{F} = 4.17\times10^{-3}\,\mathrm{F} = 4,170\mu\mathrm{F}$$

In practice a 4,700μF component would be suitable with a working voltage of at least 16V.

Electrolytic capacitors

As you've seen, the vast majority of capacitors used in power supplies are electrolytic types that permit relatively large values of capacitance in a small volume. In the manufacture of electrolytic capacitors, an oxide film (the dielectric) is grown onto a metal anode by electrochemical means. The oxide film is very thin and offers a dielectric constant of up to about 25. In normal operation, the dielectric must be polarised by the application of a DC voltage.

Electrolytic capacitors need to be treated with care; they are not as widely applicable as other types of capacitor for a number of reasons, including the need for a polarising voltage, wide tolerance, poor temperature characteristics and relatively large values of leakage current, significant effective series resistance (ESR) and effective series inductance (ESL). Electrolytic capacitors are also significantly less reliable than other types and are prone to failure at high temperatures (see later). This serves to emphasise the need to choose the correct type of capacitor for power supply applications and particularly where reliability is paramount. The properties of various common types of electrolytic capacitor are listed in Table 6.1.

Electrolytic capacitor characteristics

It's worth spending a little more time explaining why large electrolytic capacitors need to be treated carefully. We will start by explaining some important characteristics that have a major effect on the way capacitors operate.

Dielectric

A capacitor is an energy-storing device made of two parallel conductive plates separated by an insulating (dielectric) material. When a voltage is applied across the plates, the electric field in the dielectric displaces electric charges, and thus it stores energy. It is assumed that there are no free charges in the dielectric (at least in the ideal case), and that while they are displaced, they are not free to move (as they can in a conductor).

Value and tolerance

The capacitance of an electrolytic capacitor is usually stated in μF (microfarads) and is determined by the physical properties of the capacitor (effective plate area and separation) as well as the dielectric material. Tolerance is the maximum allowable deviation from the nominal capacitance value at a specified temperature. Typical values of tolerance for electrolytic capacitors can range from –20% to +50% and so you can certainly expect some deviation from any marked value of capacitance.

Maximum working voltage

The maximum working voltage for an electrolytic capacitor is defined as the maximum DC voltage that can be applied to the capacitor continuously at the rated temperature. Note that AC and DC ratings are *not* the same and the sum of the DC and any peak AC voltages simultaneously applied to the capacitor must *never* exceed the rated DC voltage.

Equivalent series resistance

ESR is the internal resistance of the capacitor expressed as a single resistance value, modeled as connected in series with the capacitor (which is then viewed as a 'perfect' component (see later)). It is important to note that ESR is made up of the sum of the resistance loss in the dielectric (important in electrolytic components) and the resistance of the conducting path between the capacitor plates and its external connections. ESR varies slightly with frequency, usually falling to a minimum value in the range associated with most switched-mode

Table 6.1 Comparison of electrolytic capacitor types

Type of capacitor	Typical range of values	Typical working voltage	Temperature range	Typical ripple current	Typical leakage current	Typical ESR	Life expectancy (hours)
Miniature radial lead electrolytic	1μF to 2,200μF	6.3V to 25V	–55°C to +105°C	0.4A to 9.7A	0.01CV or 4μA	15Ω to 350Ω	2,000 at 105°C
Miniature SMD electrolytic	1μF to 1,000μF	6.3V to 50V	–45°C to +85°C	0.01A to 0.33A	0.01CV or 3μA	0.4Ω to 166Ω	2,000 at 85°C
PCB-mounting electrolytic	220μF to 22,000μF	16V to 400V	–40°C to +85°C	0.7A to 4.1A	0.02CV	0.03Ω to 4Ω	2,000 at 85°C
SMD low-ESR electrolytic	100μF to 1,000μF	6.3V to 50V	–55°C to +105°C	0.3A to 0.45A	0.01CV or 3μA	0.17Ω to 0.4Ω	2,000 at 85°C
Low-ESR radial lead electrolytic	2.2μF to 1,000μF	16V to 100V	–55°C to +105°C	0.03A to 1.5A	0.01CV or 2μA	40mΩ to 2.8Ω	5,000 at 85°C
High-grade electrolytic	1,000μF to 47,000μF	25V to 385V	–40°C to +85°C	5A to 42A	0.006CV or 4μA	5mΩ to 70mΩ	4,000 at 85°C

power supplies (ie, 50kHz to 150kHz). Note also that the power loss in a capacitor increases with ESR. Because ESR is so important to the correct operation of a capacitor (it is often the reason why an apparently 'good' component fails to work) we shall examine this in much greater detail later.

Dissipation factor

The dissipation factor (D or DF) of a capacitor is the ratio of the effective series resistance (ESR) of a capacitor to its reactance (X_C) at a specified frequency. Dissipation factor is sometimes also referred to as 'tanδ', i.e. the tangent of the angle in which ESR (R_E) and X_C are the adjacent sides of a right-angle triangle.

$$D = \tan\delta$$

Note that dissipation factor is usually quoted for sinusoidal AC power applications and is less meaningful when conditions are non-sinusoidal (as in switched-mode power supplies). Dissipation factor is given by:

$$DF = \frac{R_E}{X_C} = \frac{R_E}{\left(\frac{1}{2\pi fC}\right)} = 2\pi fCR_E$$

To help put this into context, consider the following example. A capacitor of 68µF with an ESR of 1.5Ω is used at a frequency of 50Hz. The dissipation factor can be determined from:

$$DF = 2\pi fCR_E = 6.28 \times 50 \times 68 \times 10^{-6} \times 1.5$$
$$= 0.032 \text{ or } 3.2\%$$

Quality factor

The quality factor (Q or Q_F) is the ratio of capacitive reactance (X_C) to ESR (R_E) at a specified frequency. Q_F is the inverse of the dissipation factor, hence:

$$Q = \frac{1}{D} \text{ and } D = \frac{1}{Q}$$

Volumetric efficiency and energy density

The volumetric efficiency (VE) of a capacitor is a measure of how much capacitance is provided for a given component volume. Volumetric efficiency is mainly determined by the dielectric constant (*k*) and the construction of the capacitor. It's worth noting that for a given dielectric material, the volume of a capacitor varies roughly with the square of its maximum voltage rating. Thus a 63V capacitor is likely to have around 100 times the volume of a similar component rated at a mere 6.3V!

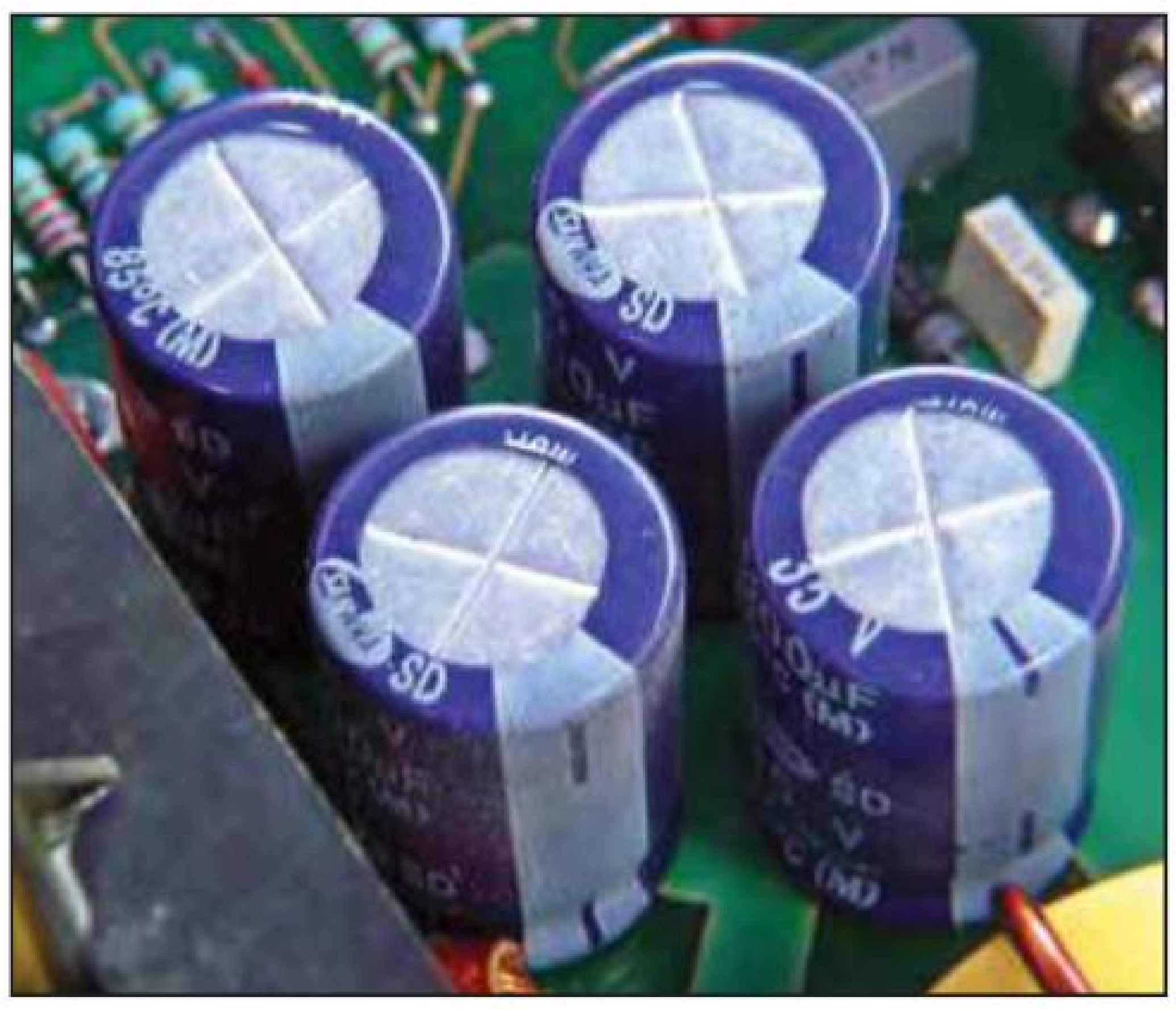


Fig.6.6. Four large electrolytic capacitors connected in parallel to form a large reservoir capacitance at the high-voltage DC input of a high-current switched-mode power supply.

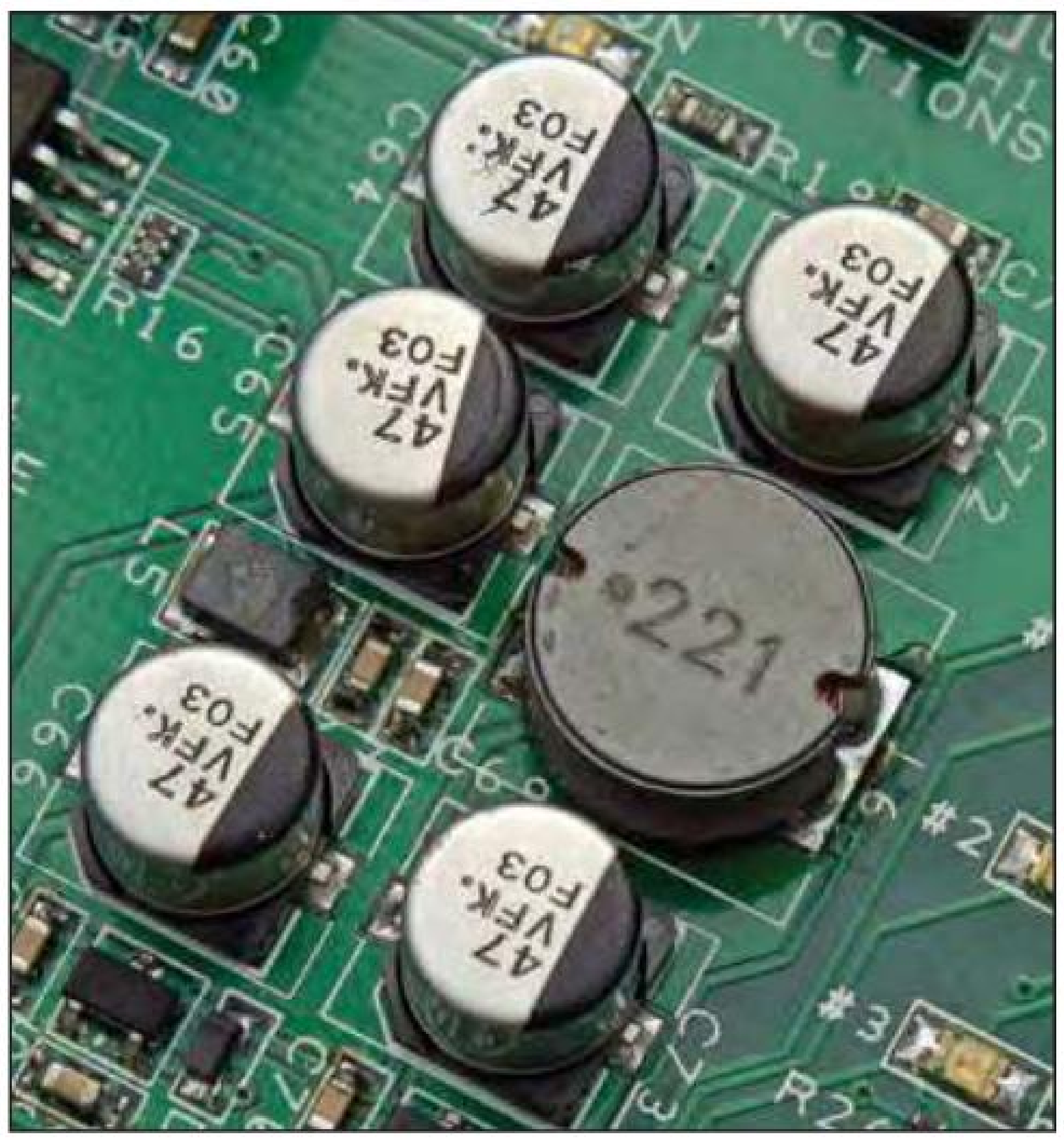


Fig.6.7. Surface-mounte reservoir capacitors in a low-power switched-mode power supply.

The energy density (ED) of a capacitor is related to its volumetric efficiency. Energy density is given by:

$$ED = \frac{C \times V}{volume}$$

where C is the capacitance and V is the working voltage.

Maximum ripple current

The maximum ripple current of a capacitor is the maximum RMS current (usually quoted in amps) that can be allowed to flow in the capacitor at a specified frequency. Note that the permissible ripple current for a capacitor is limited by the capacitor's internal temperature, which increases due to internal power dissipation associated with its ESR.

Dielectric absorption (DA)

Few electronic enthusiasts can fail to have noticed that, once charged, a capacitor seems to retain some of the charge, even after a concerted attempt is made to discharge it! Even shorting its terminals for several seconds can stubbornly fail to remove the charge from a large value capacitor. The reason for this puzzling phenomenon is dielectric absorption (sometimes referred to as 'voltage retention' or 'soakage') resulting in the generation of a potential between the terminals of a capacitor *after* it has been discharged.

Dielectric absorption can be checked using the following simple procedure:

1. Charge the capacitor to its rated voltage (eg, 63V, 100V or 150V) for five minutes.
2. Disconnect the capacitor from its charging supply.
3. Short the capacitor's terminals for five seconds, then leave the capacitor to recover for 60 seconds.
4. Connect a high-resistance voltmeter to the capacitor terminals and readthe recovery voltage present.

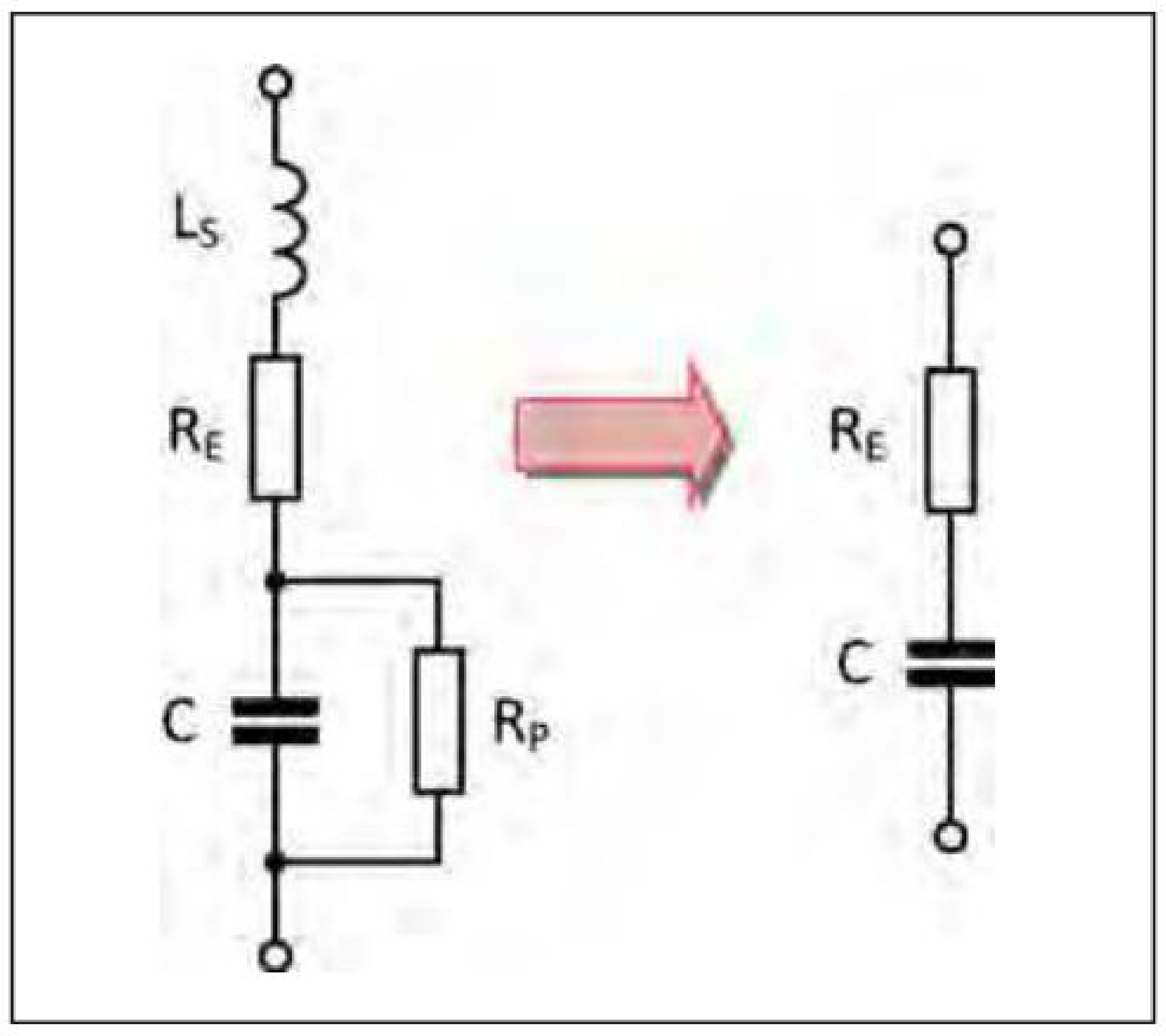


Fig.6.8. Equivalent circuit of a capacitor.

The value of dielectric absorption can be determined from:

$$DA = \frac{V_R}{V_S} \times 100\%$$

V_R is the recovery voltage and V_S is the charging supply voltage.

For example, if a capacitor is charged from a 63V supply (its rated working voltage) and it exhibits a recovery voltage of 9V, then the value of dielectric absorption will be given by:

$$DA = \frac{7}{63} \times 100\% = \frac{1}{9} \times 100\%$$
$$= 0.11 \times 100\% = 11\%$$

Equivalent series inductance

A capacitor's equivalent series inductance (ESL) is the effective inductance of the capacitor expressed as a single inductance considered to be connected in series with a perfect inductor (see Fig.6.8). It is important to note that ESL is made up of the sum of the internal inductance and the inductance of the conducting path between the capacitor plates and its external connections. ESL reduces the effectiveness of a capacitor at high frequencies. It is also responsible for a sharp dip in impedance that occurs at the series resonant frequency of some types of capacitor. Depending on the component type and value, this resonant effect occurs at frequencies of between about 700kHz for a small axial lead electrolytic capacitor to around 40MHz for small PCB-mounting film dielectric capacitors. ESL is mostly of concern for power-supply filtering in high-frequency switched-mode power supplies.

Insulation resistance

A capacitor's insulation resistance (IR) is the parallel (or 'shunt') resistance of a capacitor (see Fig.6.8). Insulation resistance is usually specified in MΩ (a typical value for an electrolytic capacitor being in the range 1MΩ to 10MΩ). The lower values of insulation resistance are associated with higher values of capacitance. Some manufacturers quote insulation resistance as a product of the insulation resistance (in MΩ) and capacitance (in µF). The insulation resistance may thus be specified in MΩ-µF.

To determine the insulation resistance for a given capacitor it is simply a matter of dividing the MΩ-µF value by the component's capacitance. For example, if a 2.2µF capacitor is taken from a range with a quoted IR value of 4.4MΩ-µF, it will have an insulation resistance of 2MΩ.

Leakage current

Leakage current is often specified in terms of the amount of stored charge (ie, as a constant multiple of capacitance (C) and applied voltage, V – typically $0.01CV$ – or as a maximum current in µA (whichever is the greater). Leakage current decreases with insulation but increases with applied voltage and temperature. Typical leakage currents for electrolytic capacitors are in the range 1µA to 10µA, and this is usually not significant in power supply applications.

Equivalent circuit of a capacitor

Having explained some of the characteristics of electrolytic capacitors you will now realise that such components are inherently imperfect and their operation is affected by internal factors such as the presence of small amounts of series resistance and inductance. We can model a real capacitor using the equivalent circuit of a capacitor shown in Fig.6.8. The components shown are:

- Effective capacitance, C
- Parallel (or 'shunt') resistance (R_P) through which a small leakage current flows
- Equivalent series resistance (ESR), R_E
- Effective series inductance (ESL), L_S.

It is important to understand that the components shown in Fig.6.8 have quite different effects on a capacitor's performance in a working circuit. For example, at low frequencies, ESL (L_S) is insignificant, it does become increasingly important at high and very high frequencies. R_P, on the other hand is of little consequence in low-impedance equipment (such as in a power supply) but it does become important in high-impedance circuit applications (for example, in a sample and hold circuit). Conversely, while R_E is unimportant in a high-impedance circuit, this component becomes critical in low-impedance situations (such as conventional and switched-mode power supplies).

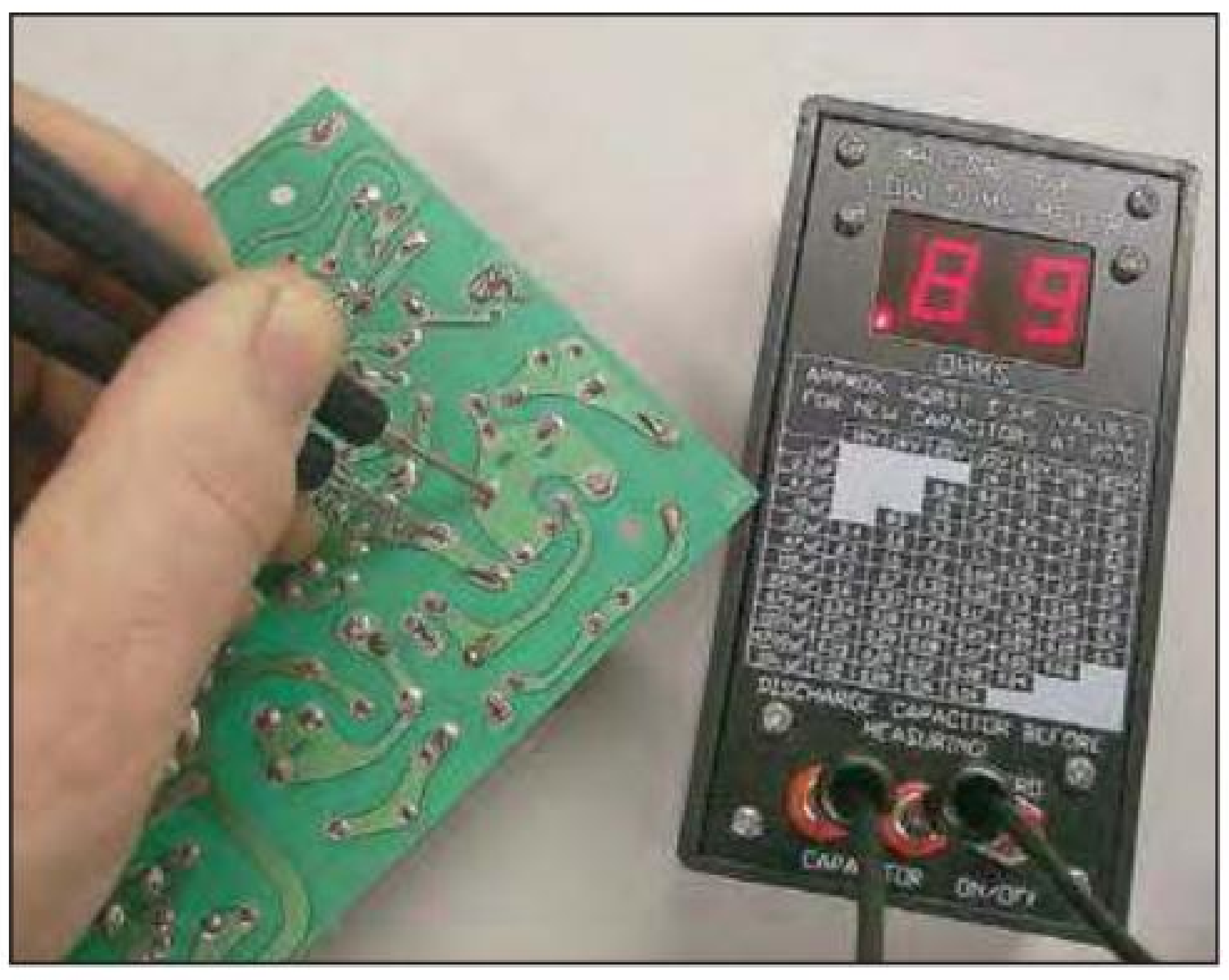


Fig.6.9 In-circuit measurement of the ESR of a PCB mounted radial lead electrolytic capacitor.

Fortunately, for most power supply applications we can simply ignore the shunt resistance and series inductance leaving us with just the ESR (R_E) to consider. Modern high-current switch-mode power supplies have increased the demand for low-impedance, high-frequency capacitors with high ripple current ratings. These components *must always* have a low value of ESR.

Capacitor failure

Capacitor failure may be either drastic or progressive. There are two types of drastic failure: short-circuit and open-circuit. The likely consequence of either condition is that the equipment to which the component is fitted will fail to operate totally or will fail to operate within specification. A progressive failure, on the other hand, is when one or more aspects of the capacitor fall outside their specified limits. For example, when a capacitor's leakage current increases to a value that is significantly greater than the maximum rating for the component.

Short-circuit failures normally result from failure of the dielectric due to voltage stress. Open-circuits, on the other hand, are usually attributable to mechanical failure of the joint between the foil and capacitor terminals. Such failures are very much less common than electrical failures in the dielectric.

Exposure to hydrogenated hydrocarbon solvent cleaners (which enter electrolytic capacitors through the pressure seal) can often result in chemical attacks on the aluminium foil electrodes of electrolytic capacitors. This type of failure is now less common due to the use of no-wash or water-washable fluxes while soldering. When electronic equipment requires cleaning, it is important to ensure that electrolytic capacitors are treated very sparingly, if at all!

Temperature effects

Failure of large electrolytic capacitors is often caused by relatively high values of working temperature which, in turn, are usually caused by relatively high values of ESR. Since ESR increases with temperature this can become a vicious circle!

The reduction in the working life of a capacitor working at a high temperature usually occurs because the electrolyte starts to evaporate, and its vapour passes through the end seal. The consequent loss of electrolyte *reduces* the capacitance and *increases* the ESR (since less electrolyte is present). In turn, this can result in early equipment failure where correct circuit operating conditions are no longer maintained due to ineffective power line filtering or decoupling.

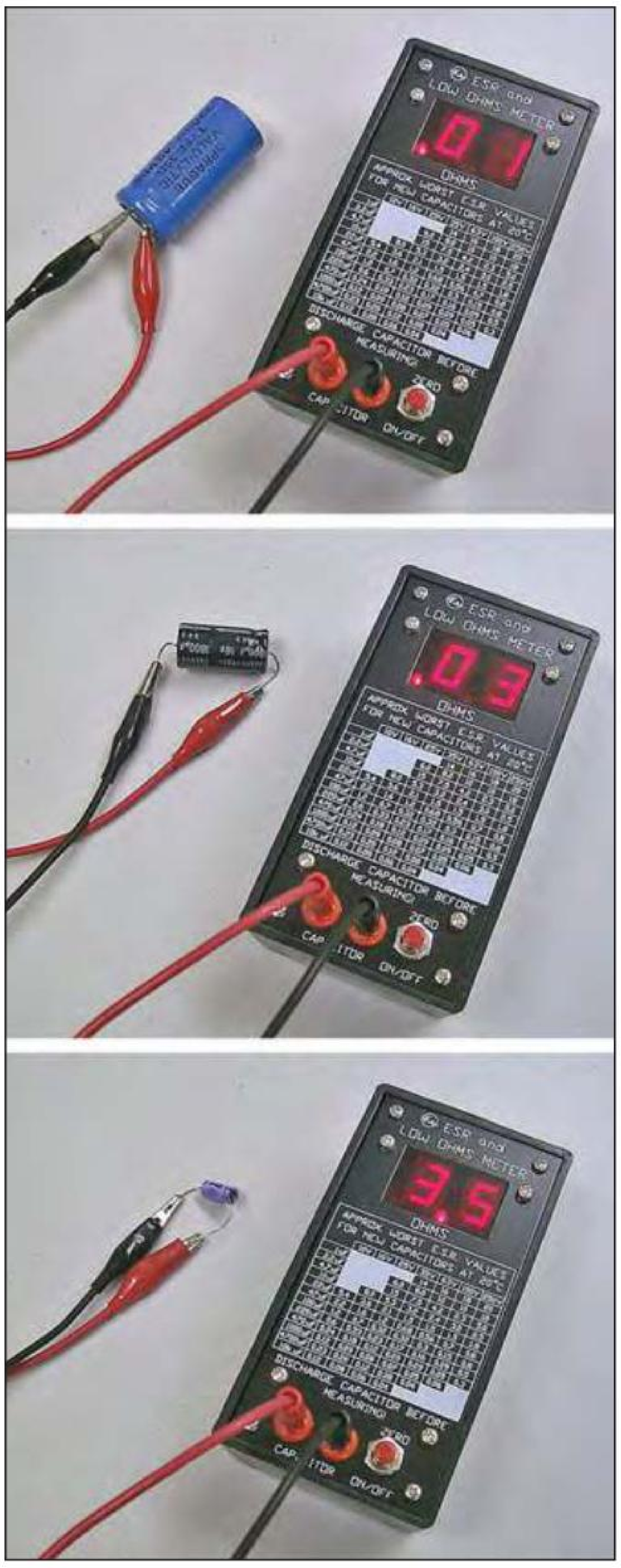

ESR and LOW OHMS METER
.01
.03
3.5
OHMS
APPROX. WORST E.S.R. VALUES FOR NEW CAPACITORS AT 20°C
DISCHARGE CAPACITOR BEFORE MEASURING
ZERO
CAPACITOR
ON/OFF

Fig.6.10. Various capacitor ESR measurements showing how ESR varies for different types and sizes of capacitor.

Another contributor to internal temperature rise is ambient temperature. The life expectancy of a capacitor is reduced significantly as the ambient temperature increases simply because this also results in evaporation of the electrolyte, which becomes worse as the temperature increases. In practice, the life expectancy of an electrolytic capacitor at 60°C is twice that when operating at 75°C. Beyond that, the life expectancy of the capacitor is halved for every 10°C increase in temperature.

It is particularly important to be aware that high internal temperatures can result in a significant increase in the capacitor's internal pressure. This situation can be relieved by several methods, including a controlled thickness vent in the can, a diaphragm in large cans, or by expulsion of the rubber end plug in miniature axial and radial lead electrolytic capacitors.

Working voltage

The voltage rating for a capacitor is related to the voltage used while forming the dielectric during manufacture and normally represents the maker's

Table 6.2 Typical characteristics for conventional electrolytic capacitors

Voltage rating (V)	10	16	25	35	50	63
Dissipation factor (D)	0.19	0.16	0.14	0.12	0.1	0.1
ESR (Ω)	1.15	0.97	0.84	0.72	0.6	0.6
Max ripple current (mA)	260	320	430	480	540	580

Table 6.3 Typical characteristics for 'low-ESR' electrolytic capacitors

Voltage rating (V)	10	16	25	35	50	63
Dissipation factor (D)	0.15	0.1	0.08	0.07	0.06	0.05
ESR (Ω)	0.5	0.4	0.32	0.28	0.22	0.21
Max. ripple current (mA)	380	450	600	700	900	1,100

recommended maximum value. With older electrolytic capacitors it was possible for the dielectric to slowly deteriorate when the capacitor was operated at significantly less than its rated value. Nowadays, use of high purity aluminium foil has largely eliminated this effect. Indeed, the effect of using an electrolytic capacitor at less than its rated value is to increase the reliability of the dielectric in accordance with a fifth power law. It's worth illustrating this with two contrasting examples:

Capacitor A

Capacitor value	10,000µF
Manufacturer's voltage rating	100V
Actual working (operating) voltage	67V
Basic failure rate	0.1% / 1,000 hrs

Capacitor B

Capacitor value	6,800µF
Manufacturer's voltage rating	50V
Actual working (operating) voltage	50V
Basic failure rate	0.05% per 1,000 hrs

The effective failure rate (EFR) for a capacitor can be calculated from:

$$EFR = BFR \times \left(\frac{V_W}{V_{MR}}\right)^5 \text{ \% per 1,000 hours}$$

where BFR is the basic failure rate (% per 1,000 hours), V_W is the actual working (operating) voltage (in volts), and V_{MR} is the manufacturer's rated voltage.

Hence, the effective failure rate for capacitor A will be:

$$EFR = 0.1 \times \left(\frac{67}{100}\right)^5 = 0.1 \times 0.135$$

$$= 0.0135\text{\% per 1000 hours}$$

While the effective failure at for capacitor B will be:

$$EFR = 0.05 \times \left(\frac{100}{100}\right)^5 = 0.05 \times 1$$

$$= 0.05\text{\% per 1000 hours}$$

From this it is worth noting that, even with a lower basic failure rate, capacitor B is over three times more likely to fail than capacitor A!

Dissipation factor and ESR

Table 6.2 lists typical electrolytic capacitor characteristics in relation to voltage rating. Note how dissipation factor decreases as the voltage rating increases. Higher-voltage electrolytics tend to have a larger case sizes together with lower ESR values (see Table 6.2). As a result, internal temperature and dissipation factor can be significantly reduced by selecting a component with a higher voltage rating than is strictly necessary for a given application. By limiting internal dissipation, the working temperature of the capacitor is reduced, enhancing dielectric life and ensuring long term reliability. Thus, when the time comes to select an electrolytic component for use as a reservoir capacitor or ripple filter, it is generally wise to apply some voltage de-rating.

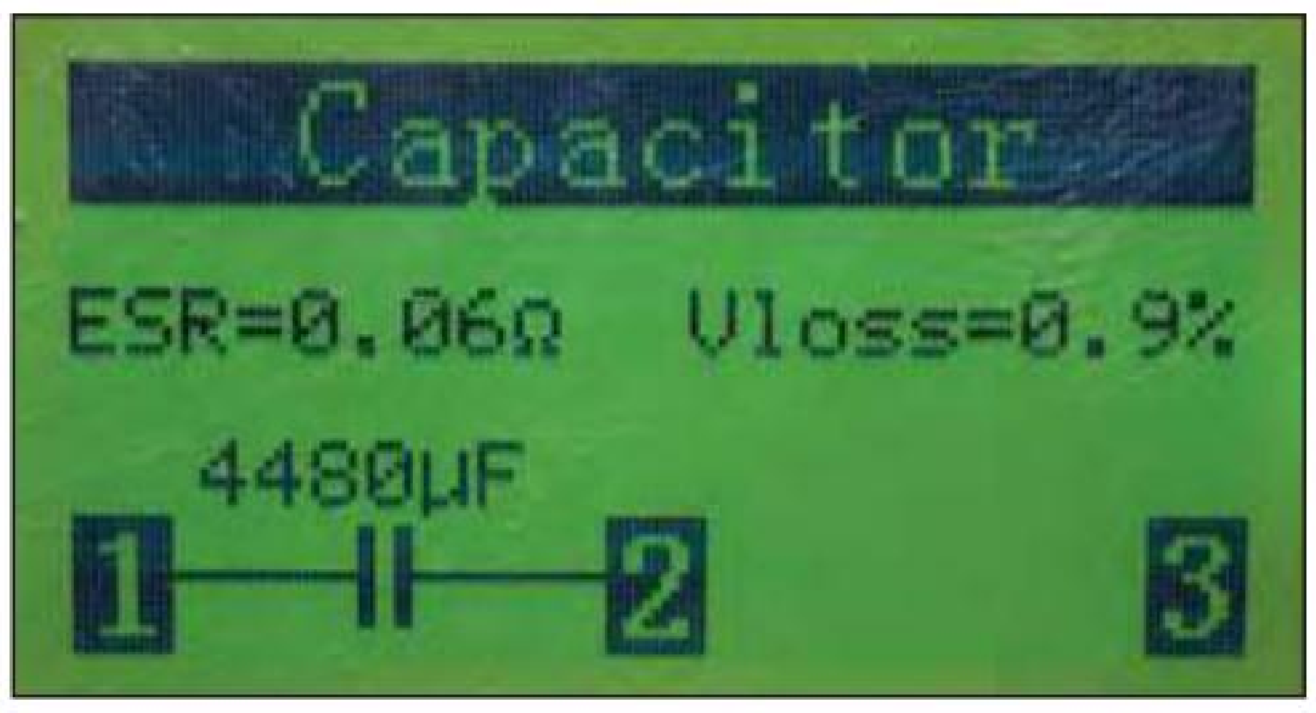


Fig.6.11. Some low-cost component checkers incorporate capacitor ESR measurement. Here a good quality 4700µF axial lead capacitor is being tested.

Practical Project: 9V-to-15V CMOS Logic Supply Converter

This month's *Practical Project* takes the form of a 9V-to-15V voltage converter based on a low-cost 555 timer and voltage doubler. This circuit is ideal for supplying power to CMOS logic devices that require a nominal 15V supply. The specifications are shown in Table 6.4.

Table 6.4 Specifications for the 9V-to-15V CMOS Logic Supply Converter

Input voltage	9V
Output voltage	15V (±0.5V)
Load current	15mA nominal (25mA max)
Switching frequency	32kHz approx
Load regulation	13% approx
Output resistance	100Ω approx

The circuit of our *9V-to-15V CMOS Logic Supply Converter* is shown in Fig.6.12.

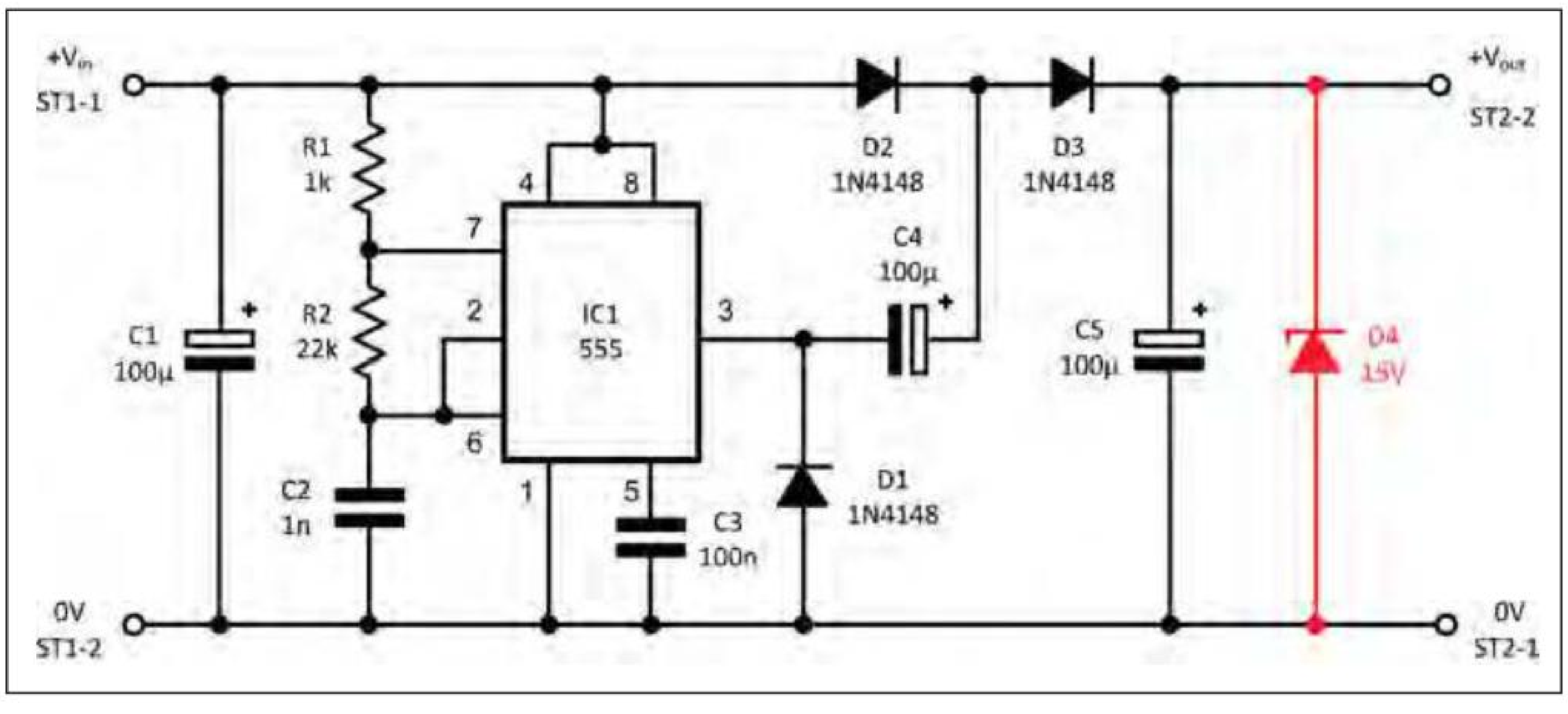


Fig.6.12. Teach-In Practical Project: 9V-to-15V CMOS Logic Supply Converter. (To improve output voltage accuracy, a 15V Zener diode (in red) can be added.)

A 555 timer (IC1) is used as an astable oscillator with its operating frequency determined by R1, R2 and C2. The square wave output from pin-3 at approximately 32kHz is fed to a simple voltage-doubler arrangement comprising D2/D3 and C4/C5. The final DC output appearing across C5 is approximately double that of the supply (less the forward voltage drops associated with D2 and D3).

You will need...

Perforated copper stripboard (9 strips each with 25 holes)
2 2-way PCB screw terminal connector (ST1 and ST2)
1 1kΩ resistor (R1)
1 22kΩ resistor (R2)
3 100µF 35V capacitors (C1, C4 and C5)
1 1nF ceramic capacitor (C2)
1 100nF ceramic capacitor (C3)
1 555 8-pin DIL timer (IC1)
1 low-profile 8-pin DIL socket
3 1N4148 diodes (D1 to D3)
1 15V Zener diode (optional, see text)
4 stand-off pillars and mounting screws

Construction

The stripboard layout of the 9V-to-15V CMOS logic supply converter is shown in Fig.6.13. Note that there are 23 track

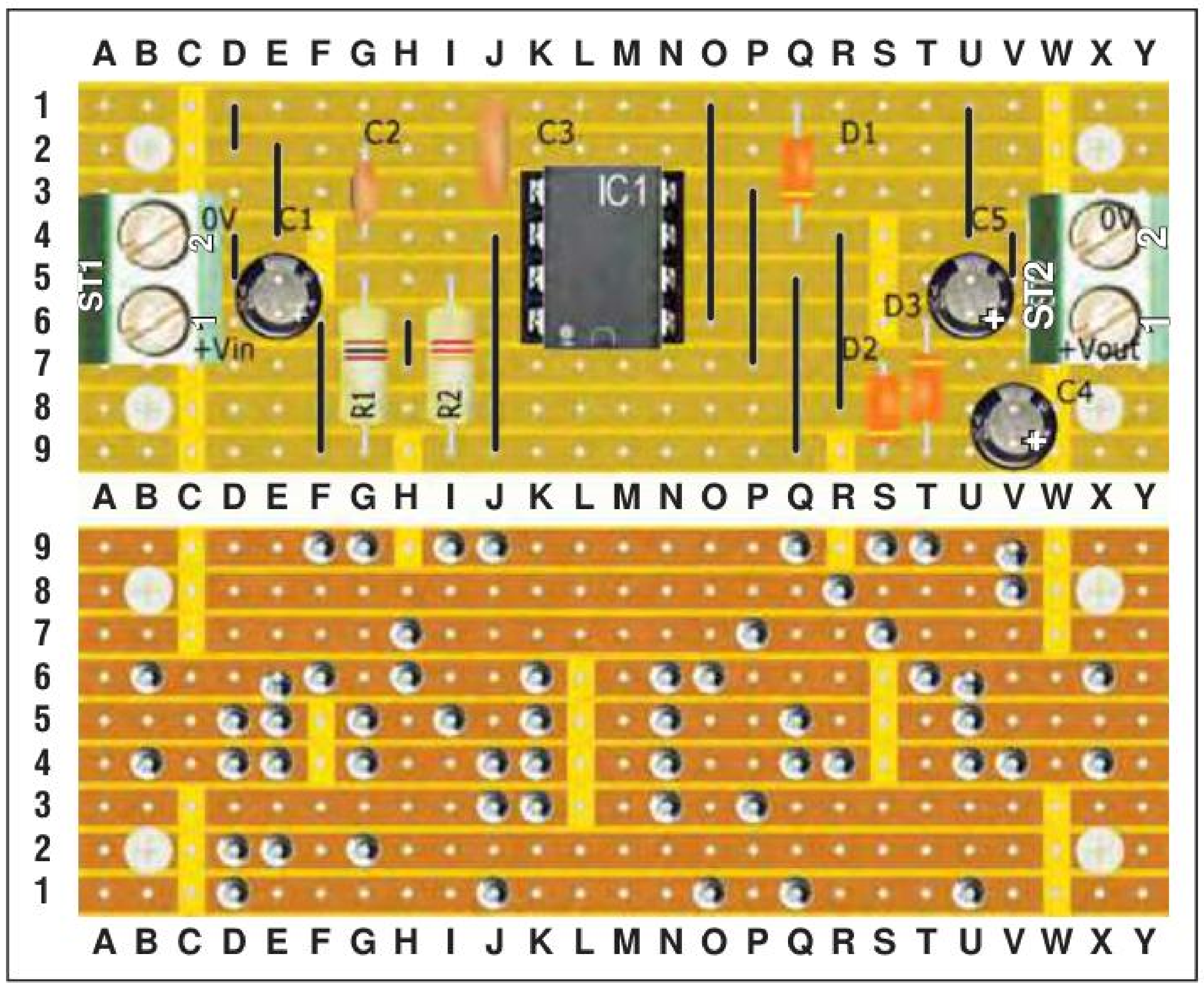


Fig.6.13. Stripboard layout of the 9V-to-15V CMOS Logic Supply Converter.

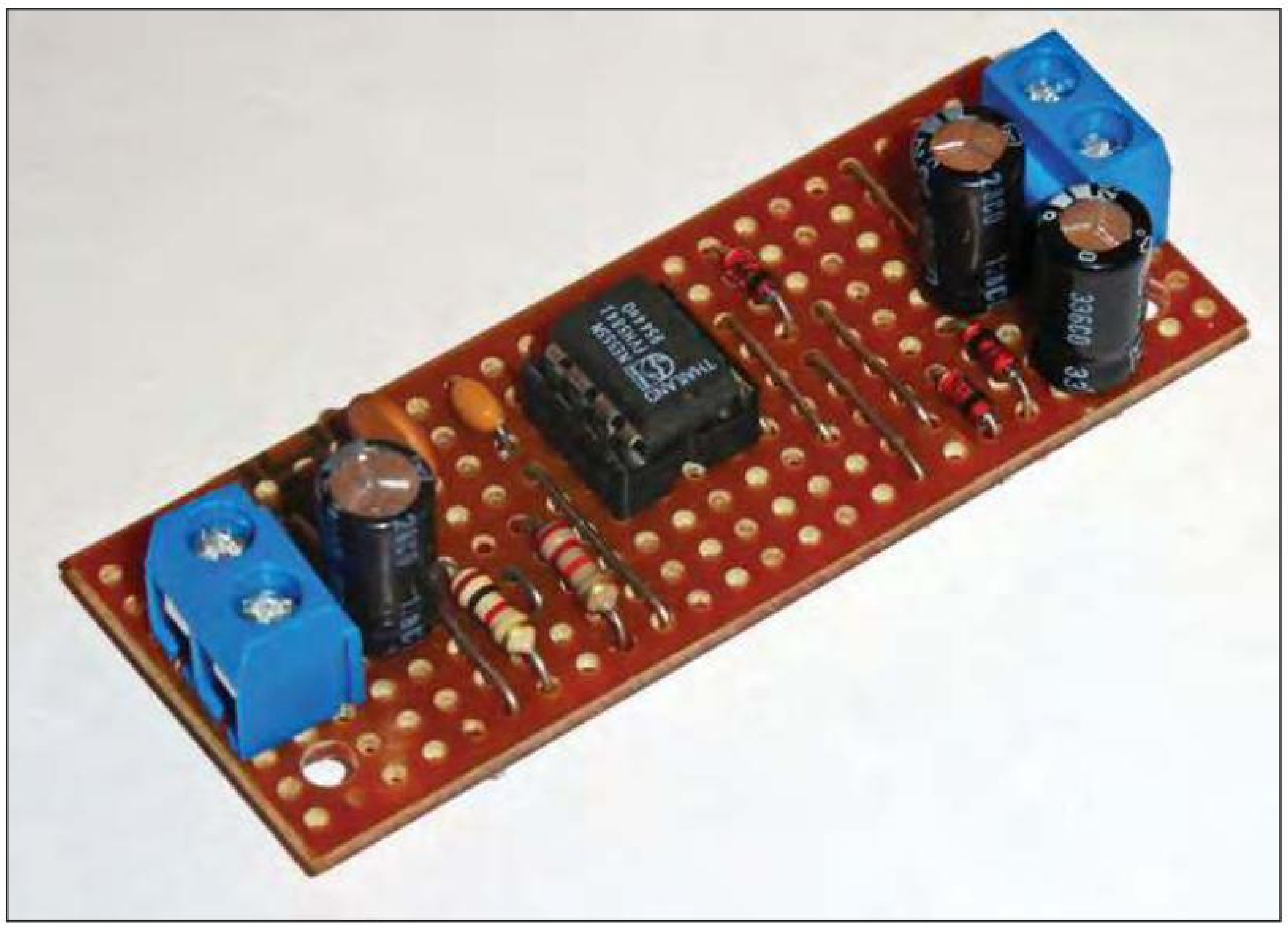


Fig.6.14. The finished 9V-to-15V CMOS Logic Supply Converter.

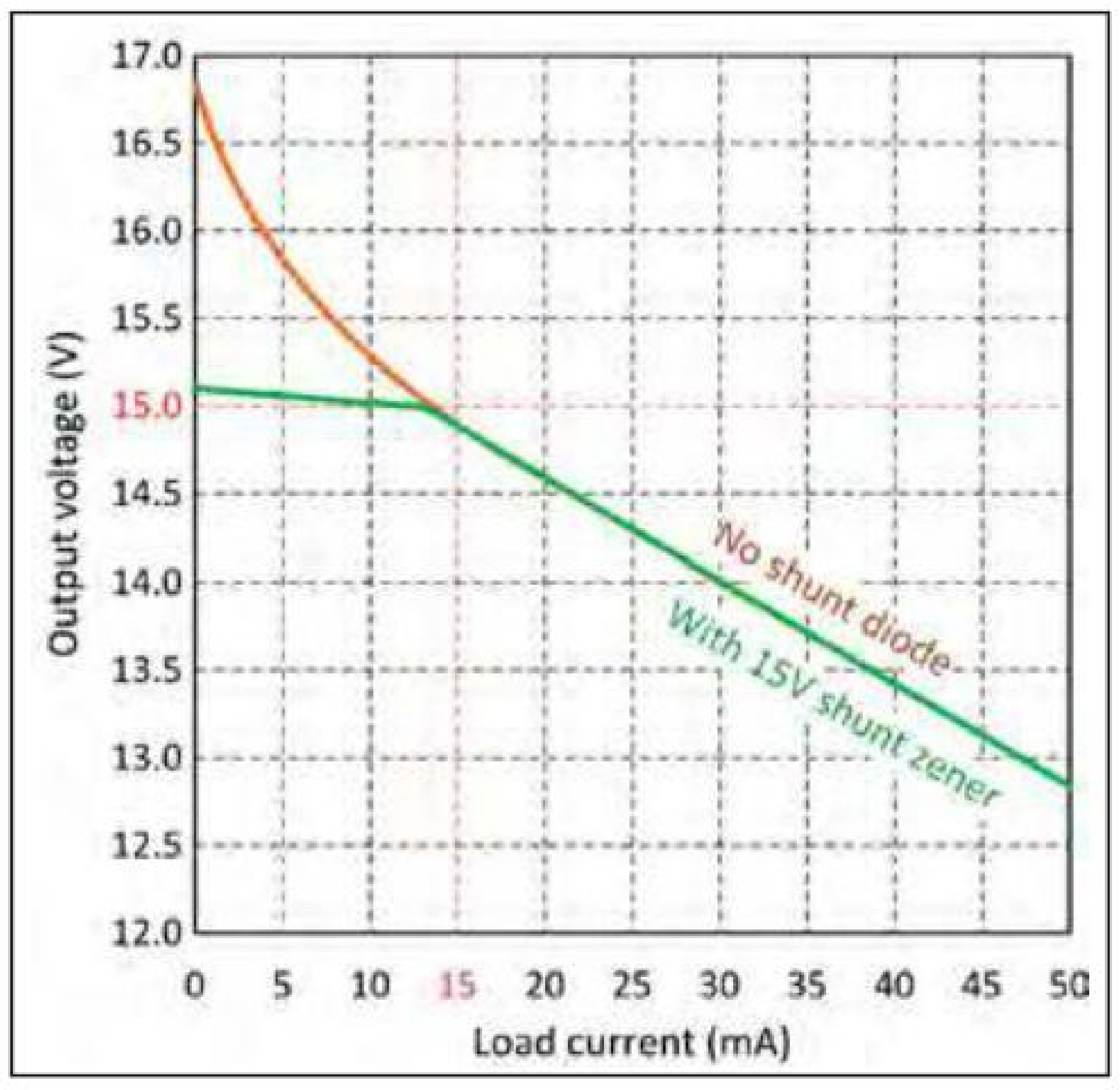
breaks on the underside of the board and 12 links on the upper side. As always, once assembly is complete it is well worth carrying out a careful inspection of the circuit board, particularly checking the off-board wiring and links to ST1 and ST2.

The off-load output voltage of the 9V-to-15V CMOS supply converter is just less than 17V. Due to its inherent ability to operate over a wide range of supply voltages, this is not usually a problem with CMOS logic. However, if it is necessary to maintain the output much closer to the nominal 15V then a 15V shunt Zener diode can simply be connected in parallel with C5, as shown in red in Fig.6.12. The off-load output voltage will then be nearer to 15.1V, falling to 14.9V at the nominal 15mA load current (see Fig.6.15).

Next month

In next month's *Teach-In 2019* we will be introducing negative voltage converters and taking a detailed look at thermal design considerations for power supplies. Our *Practical Project* will take the form of a simple low-current +12V to –9V converter.

Fig.6.15. Load regulation characteristic for the 9V-to-15V CMOS Logic Supply Converter.



Get it Right! – Selecting and using reservoir capacitors

Failure can be catastrophic, so electrolytic capacitors should always be selected and used with care. The following precautions are recommended:

1. Always ensure that a capacitor is correctly rated in terms of value, working voltage, ripple current, dissipation factor and ESR.

2. Never exceed the working voltage and ripple current ratings of a capacitor. There is a very risk of overheating and explosion if electrolytic components are operated beyond their ratings!

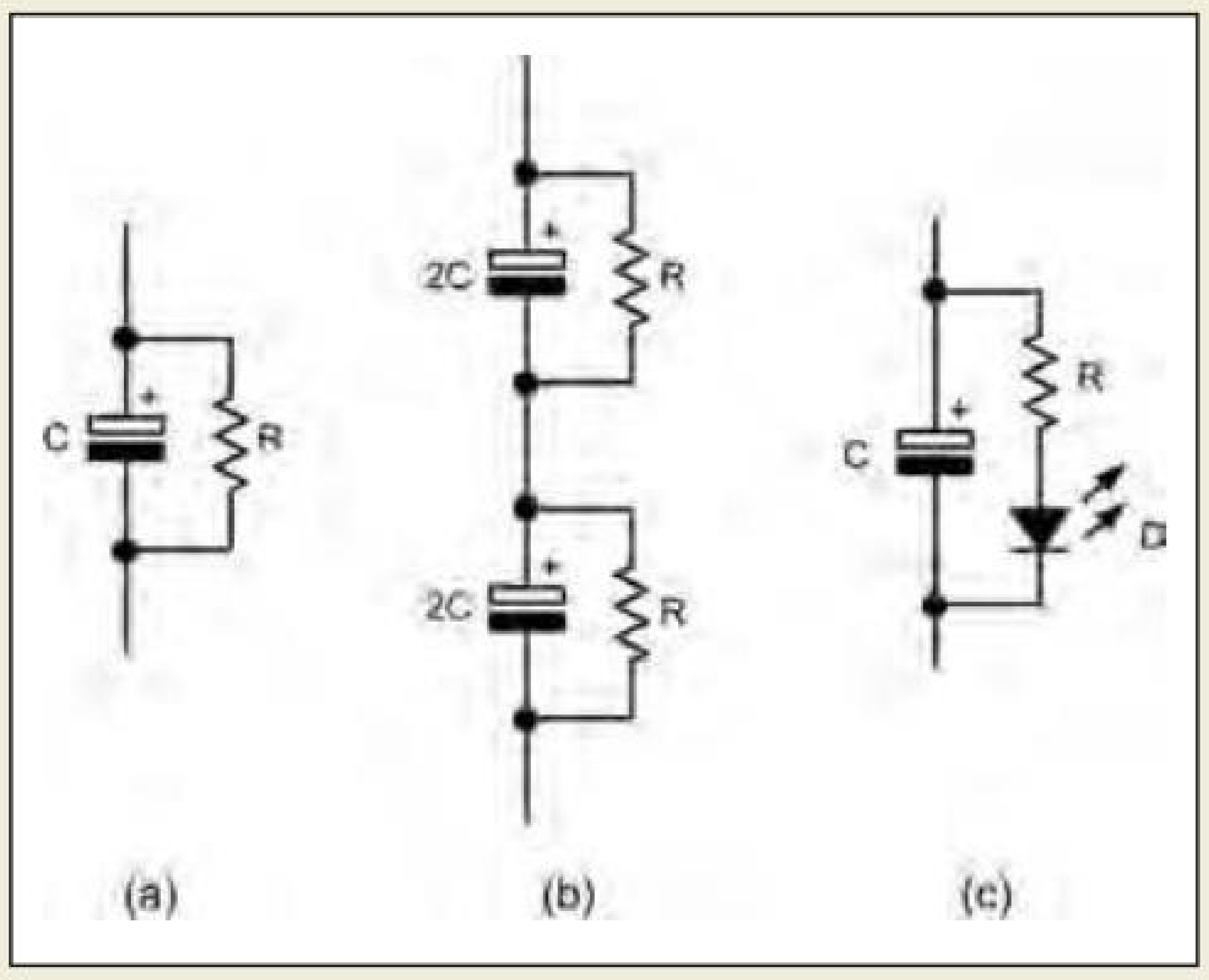


Fig.6.16. Reservoir capacitor arrangements.

3. In critical applications, always use 'low-ESR' capacitors. If in doubt, use a larger value of capacitor with a lower value of ESR in order to improve performance and reduce internal temperature rise.

4. Remember, ESR increases with temperature, so power supply cooling is important. In order to permit convection cooling check that there is sufficient space around large capacitors and avoid placing them in close proximity to heatsinks and dissipators.

5. Large electrolytic capacitors can retain their charge for long periods after the supply has been disconnected. It is essential to ensure that such components are fully discharged before attempting to work on circuits. To ensure that this is the case, a high-value fixed resistor can be connected across the capacitor's terminals, as shown in Fig.6.16(a). Suitable values for resistor *R* range from around 47kΩ to 1MΩ.

6. Electrolytic capacitors that have been kept in store for long periods will invariably be less reliable than new-stock components. Where performance and reliability are important, you should avoid using surplus and old-stock components.

7. Where electrolytic capacitors are connected in series it is advisable to connect fixed voltage equalising resistors in parallel with each component, as shown in Fig.6.16(b). Suitable values for R usually range from around 10kΩ to 100kΩ, depending upon working voltage.

8. A simple LED indicator (see Fig.6.16(c) can be fitted to indicate that charge is present in large electrolytic components. This also provides a discharge path (see point 5). Suitable values for *R* range from 1kΩ to 47kΩ, depending upon working voltage.

9. Electrolytic capacitors should be disposed of with care, particularly in cases where the seal may have ruptured and chemical material may have become exposed.

AUDIO OUT

By Jake Rothman

Backing Baxandall – Part 2

Article by Jake Rothman
PCB and circuit by Mike Grindle

The circuit presented here is a very small Baxandall tone control module designed by Mike Grindle; owner of 'Life is Unfair Audio Devices' (**www.facebook.com/lifeisunfairaudio**) who make and design bespoke audio products, mainly guitar pedals. This circuit is specially designed as a little add-on for fuzz pedals and other effects, but as pointed out in *Part 1*, it can be used in inexpensive Hi-Fi. (In a future issue I'll present a more refined Hi-Fi tone control using 5532 op amps.) This module is based on the TL072/62 Bi-Fet op amp, which suits high impedance sources like electric guitar. However, you can use the 5532 op amp, but you will need to make a few small mods, which I will detail below.

Note that the input bias current for the op amp goes directly through the bass pots wiper, which could cause noise. This only matters with bipolar op amps, such as the 5532 and 4558. With Bi-Fet op amps, their bias currents are so low it is not a problem. Note also, that with bipolar op amps, the input bias resistor R2 should be dropped to 220kΩ. If DC blocking is required, put a 10µF capacitor in series with R5 and put a 220kΩ resistor across C6 to provide a DC negative feedback path (see Fig.16).

Battery power and op amp selection

Most guitar foot pedals work on a +9V power rail, rather than the usual ±12V/15V op amps normally use. You have to be careful in these situations to get the bias point optimised. Theoretically, it is +4.5V; however, most op amps find it more difficult to swing right down to 0V than up to V+. So, for the TL072 it is best to bias it at 4.9V, making R10 equal to 82kΩ (Fig.8 in *Part 1*). This will give a maximum output of 6.8V pk-pk, as opposed to 6V pk-pk with 4.5V of bias. The current consumption is 2.8mA. The low power (0.5mA) TL062 gives more voltage swing, providing 7.8V with R10 = 91kΩ. The 5532 is not really suitable for 9V because the negative cycle drive is reduced under load unless it is biased to +6V. It's much better to run a 5532 at +24V.

Stability

If you opt for the 5532 approach then note that there is a specific problem with high-frequency stability when it is used at gains below unity (as occurs in tone controls on the cut settings). This is caused by a reduced phase margin with the internal feedback compensation. One way round this is to use the single version – the 5534 – and over compensate it by pushing C10 from 22pF to 39pF (Fig.16), which will reduce slew rate. A better method is to put a series RC network from the inverting input to ground, see R5-C6 in Fig.16. This will increase noise gain, but at inaudible frequencies. Finally, long leads to the pots can cause instability, this can be fixed by increasing the feedback phase-lead capacitor (C8) to say, 470pF (max).

The component values shown in Fig.8 (*Part 1* – see last month) are optimised for guitar, with the pivot point at 800Hz rather than the normal 1kHz Hi-Fi figure. The curves are shown in Fig.17. C6 is increased from 68pF to 2.2nF on the guitar version to prevent the treble rise from going above 10kHz. The same with the output capacitor (C7), which rolls off below 50Hz, whereas for Hi-Fi it should be upped to 10µF.

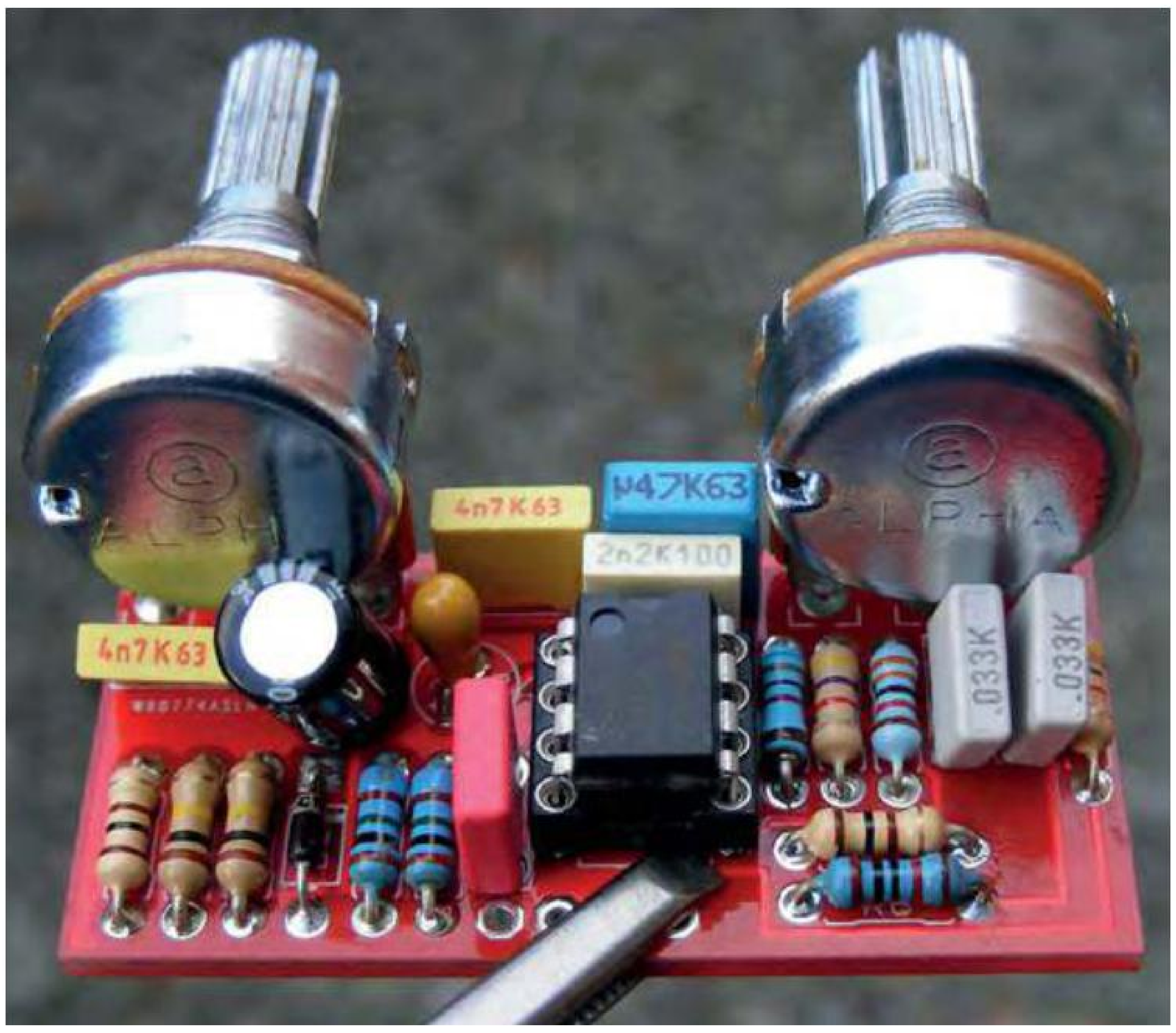


Fig.15. Mike Grindle's PCB for the Baxandall circuit given in Fig.8 in Part 1 last month.

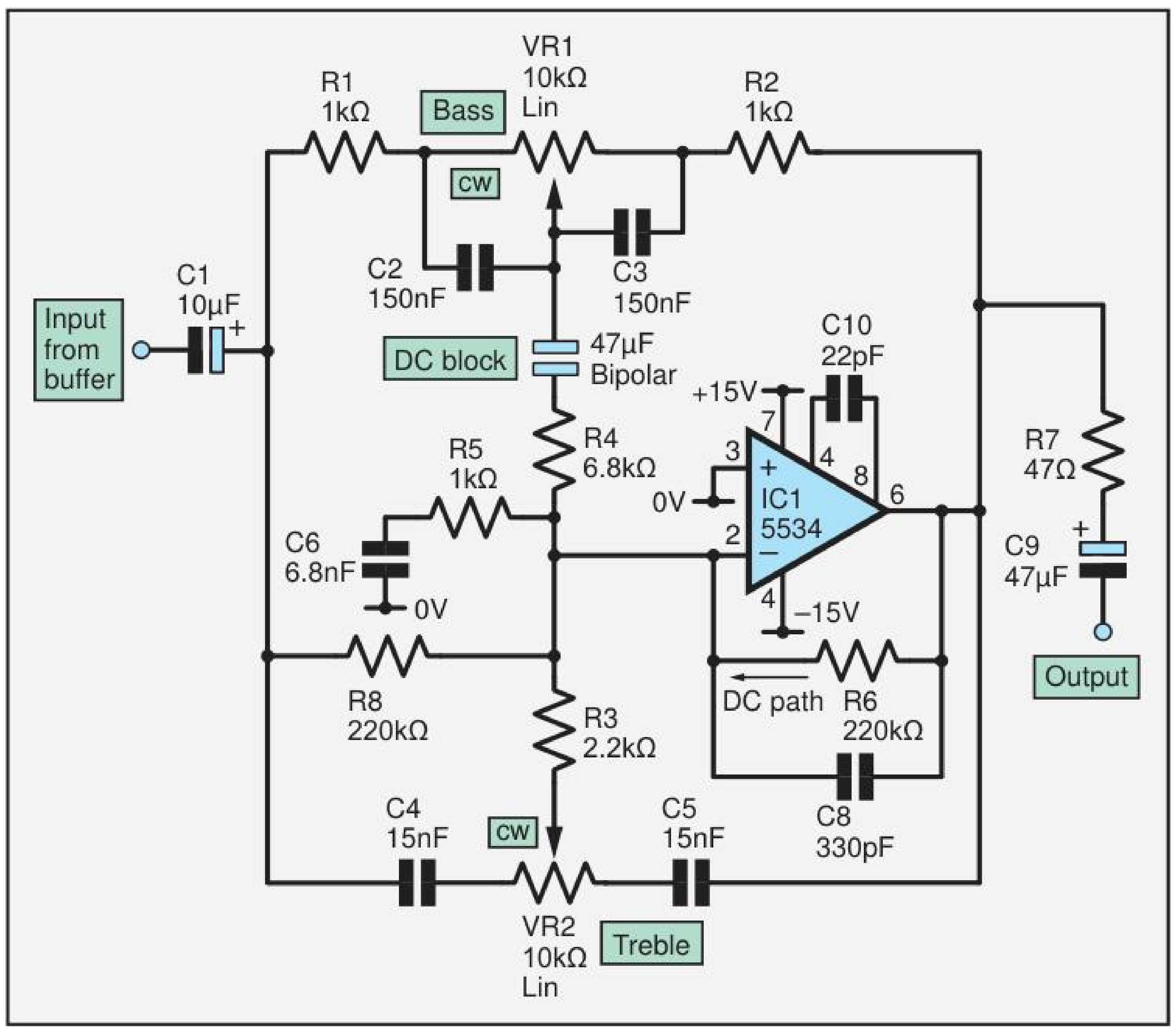


Fig.16. Baxandall circuit using 5532 op amp. Note compensation network (R5, C6), DC feedback path (R6) and lower impedances.

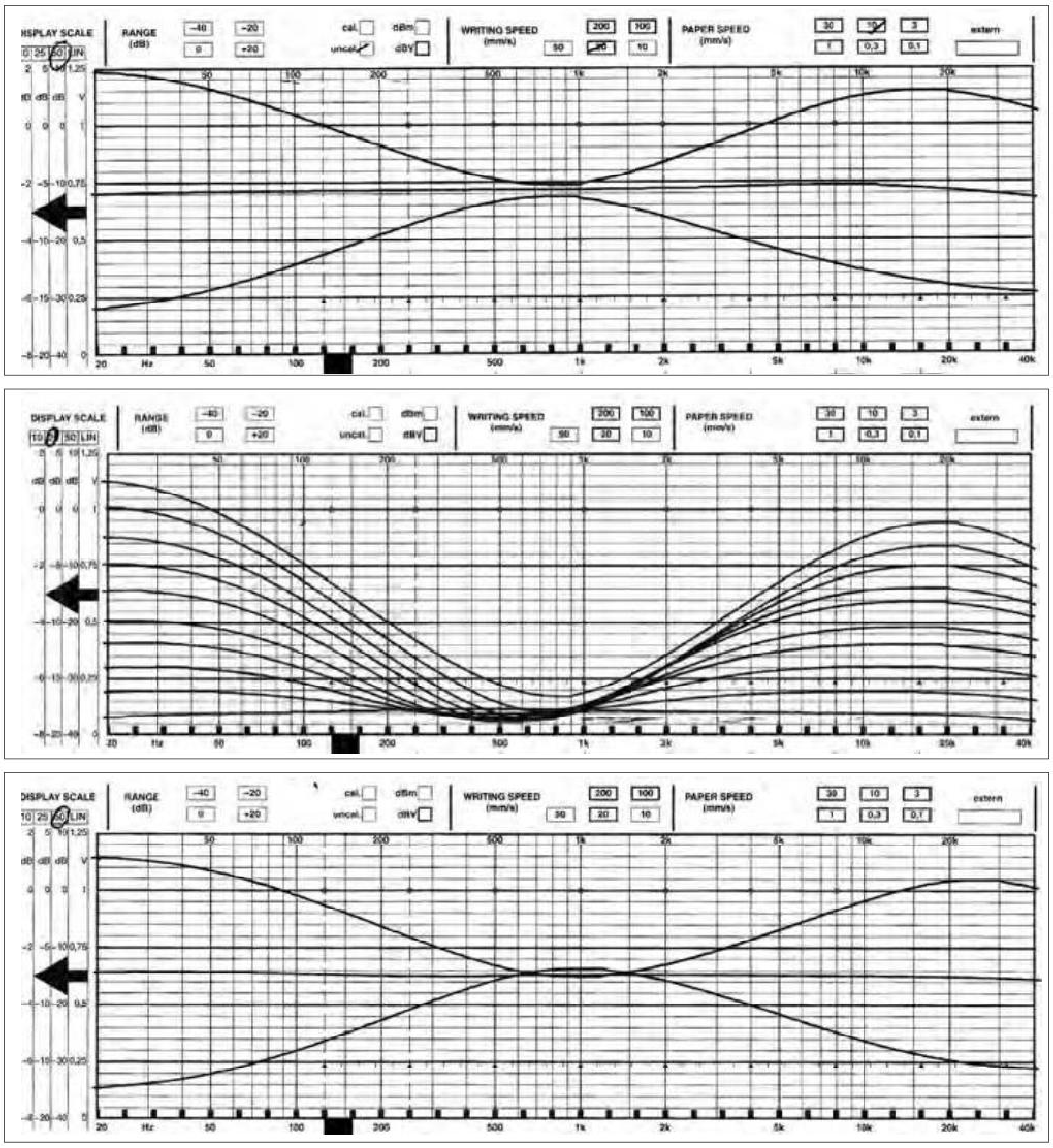


Fig.17. Frequency response curves for the Mike Grindle Baxandall tone control circuit in Fig.8 and Fig.15: (top) 800Hz version; (middle) the pots being incremented, illustrating a small amount of interaction (gain and frequency shifts) between the pots – you can clearly see the movement of the pivot point in the bass; (bottom) the 1kHz version.

There is also a little bit of power supply housekeeping, such as decoupling capacitor C8 and current-limiting resistor R9 in Fig.8 in *Part 1* of this article. Power supply reverse-protection diode D1 is also fitted because guitarists are notorious for subjecting equipment to random power supplies. This could be replaced with a Zener diode if you want to stabilise the supply at a particular voltage; for example, if the unit is to be powered off another supply, such as from a power amplifier. R9 would of course have to have the correct value for your choice of Zener diode voltage.

Construction

The PCB follows the increasingly popular method of home construction, whereby a CAD-generated file (eg, from Autodesk Eagle) is sent to a Chinese fabrication house, in this case **www.pcbway.com** (see Fig.18). On these boards, plated-through holes, component ledgending and solder resist are standard. The top and bottom track layouts and overlay are shown in Fig.19. These features are difficult to achieve with a typical home workshop using a UV light and ferric chloride etching set-up. However, for Chinese PCBs there is the problem of a turnaround of a few weeks. I must confess that despite the mess, I still like the immediacy of home etching for R&D and prototyping. Checking and modifications can be done in an hour or so in a serendipitous uninterrupted cognitive loop. Normally, when one commits one's design to a fab house it has to be right first time. Both routes have pros and cons. For this project, the relevant files are available on the *Practical Electronics* website.

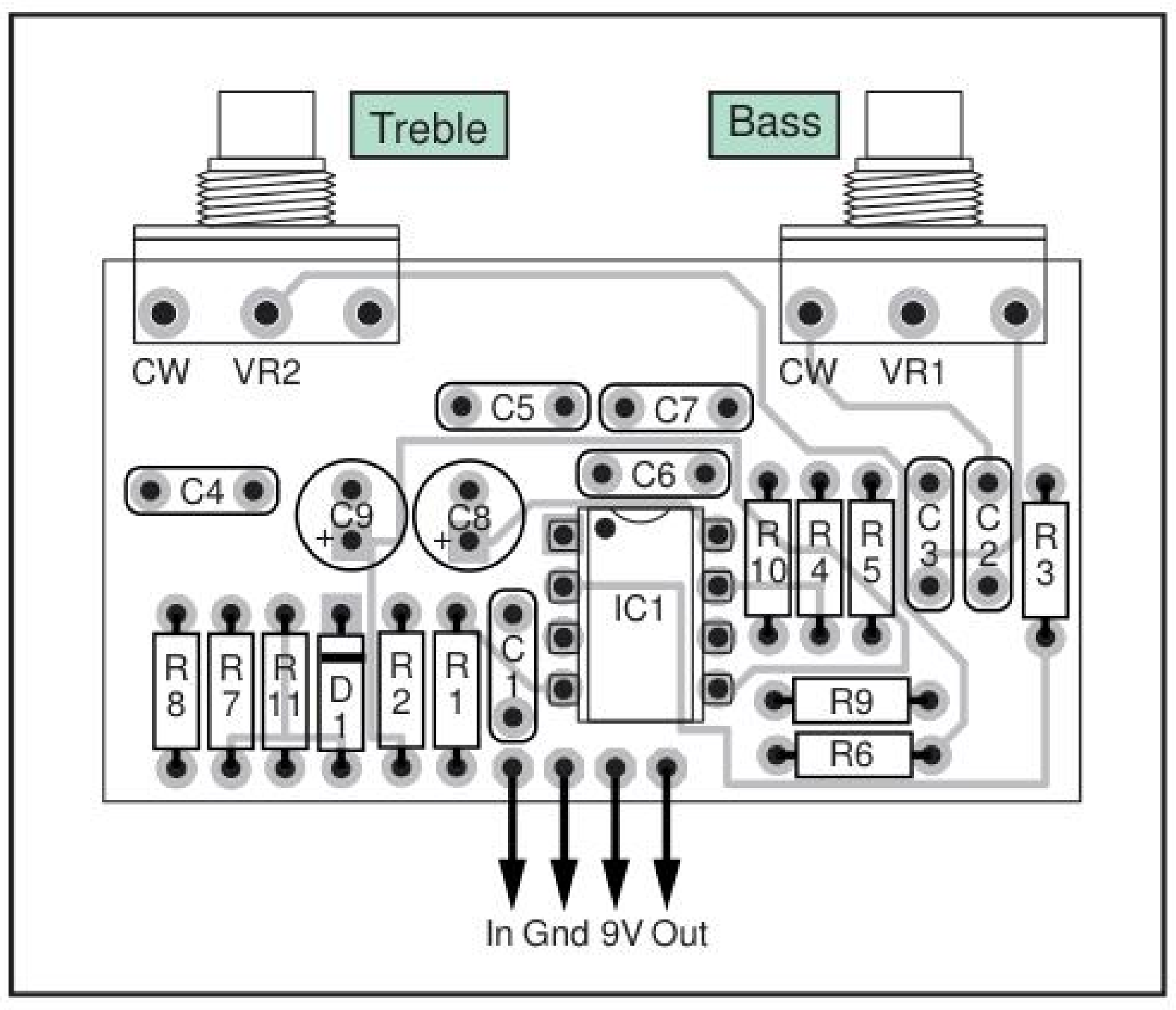


Fig.19. Component overlay for the PCB designed by Mike Grindle. Only the top layer of track is shown. The design is avilable for download from the *EPE / Practical Electronics* website.

Components

The board was originally designed for right-angle PCB-mount pots, as shown in Fig.20, which are very popular among the guitar pedal building community. With plated-through holes they are strong enough to support small PCBs directly without the need for

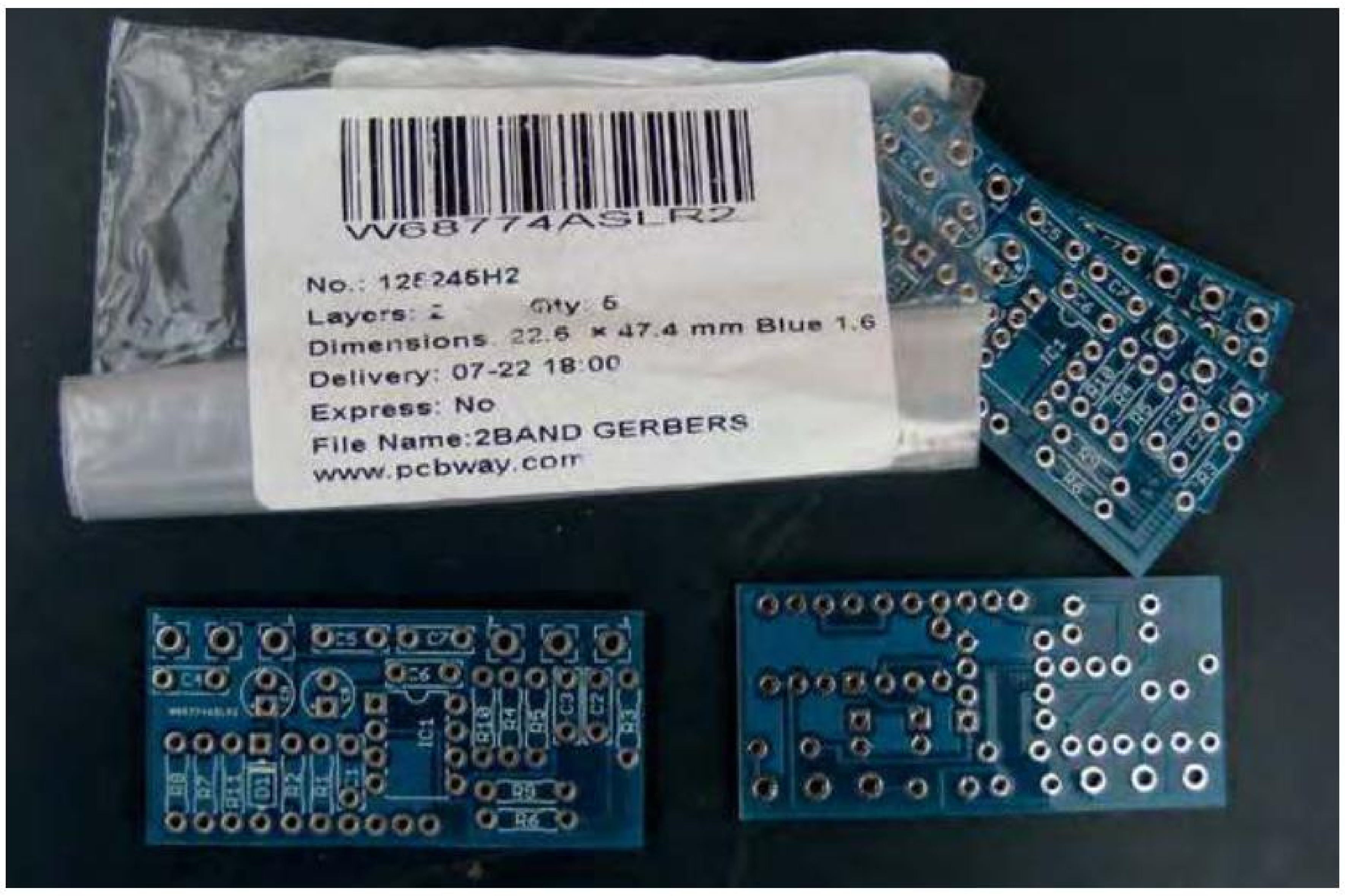


Fig.18. PCBs fresh from PCBWay.com

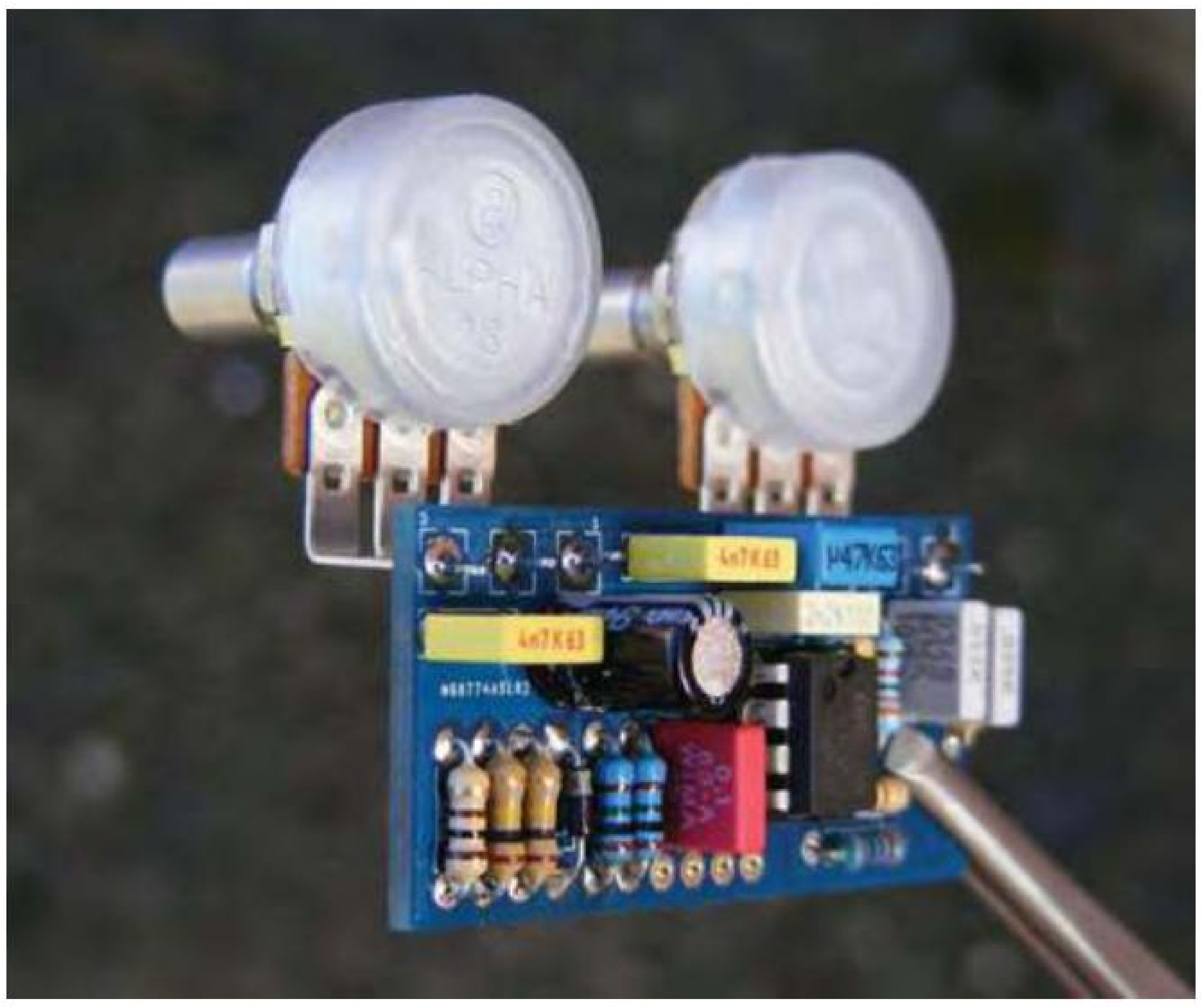


Fig.20. The board can be mounted with right-angle pots.

pillars. Note they are mounted on the underside of the board if they are to turn the right way. These are available from online supplier Tayda, which has become one of the main suppliers for electronic music constructors. If you have to use the standard vertical mount pots, capacitors C2, C3, C9 and C4 will have to be bent over a bit as shown in Fig.21. Later boards and the *PE* files for download will have the pot positions moved forwards to avoid this.

Modifications

For stereo, two PCBs will be needed with the rear gang of a pair of dual-gang pots fitted into one PCB, as shown in Fig.22. The other PCB will have to be connected with wire links. For lower noise, use the 5532 and *increase* all capacitors by a factor of five and *reduce* all resistors by a factor of five, as shown in Fig.16. This impedance reduction technique is a standard audio trick popularised by Baxandall. He advised speaker manufacturer KEF when their active speakers were hissing, 'to up the level, drop the impedance'.

Component List

Resistors (*All 5% 0.25W carbon film*)
R1, R2 1MΩ
R3, R4 4.7kΩ
R5 33kΩ
R6 10kΩ
R7 100kΩ
R8, R9 100Ω
R10, 100kΩ (or reduced to 82kΩ for TL072 or 91kΩ for TL062)
R11 100kΩ

Potentiometers
VR1, VR2 50kΩ or 47kΩ linear 16mm. Vertical, Alpha RV16AF Rapid 65-0725 or Tayda right-angle mount PCB type A-2393.

Capacitors (*All 5mm polyester*)
C1 100nF
C2, C3 33nF (47nF for 1kHz)
C4, C5 4.7nF (2.7nF for 1kHz)
C6 2.2nF (reduce to 68pF for Hi-Fi)
C7 470nF (increase to 10µF for Hi-Fi)
C8 100µF C9 4.7µF 2.5mm 10V or greater radial electrolytic

Semiconductors
TL072 or TL062 dual Bi-Fet op amp
D1 1N4001
*D1 can be Zener to suit power supply

Japanese 'Baxandall'

By putting the passive tone control circuit in the negative feedback loop of an op amp. It is possible to obtain the sonically superior *passive* curve shape with the desirable *active* gain characteristic of the normal Baxandall. It also avoids phase inversion and low input impedance. There is one major problem, anti-log pots are needed because the boost/cut direction is reversed. Getting dual-gang anti-log centre-detent pots can be difficult. A typical circuit is shown in Fig.23, similar to that used in the Sansui A80 and NAD 3020 amplifiers.

End note

Peter Baxandall was the consummate analogue audio engineer whose influence has not yet been fully appreciated. It's time his biography and technical legacy were published, although this would be a research challenge because his consultancy reports and letters are scattered among numerous engineers and companies. Much of this material resides with those who've worked at KEF, QUAD and Calrec, to name a few. Douglas Self has published some of the correspondence he had in a Linear Audio publication, *Baxandall and Self on Audio Power*, and that's it for the moment. Luckily, all of Baxandall's *Wireless World* articles can be viewed at: **americanradiohistory.com**

There are also articles he wrote for the *Journal of the Audio Engineering Society*, but these can only be accessed via the AES Electronic Library, for which you'd need to pay. Alternatively, you could join the AES – Peter Baxandall was a lifelong member.

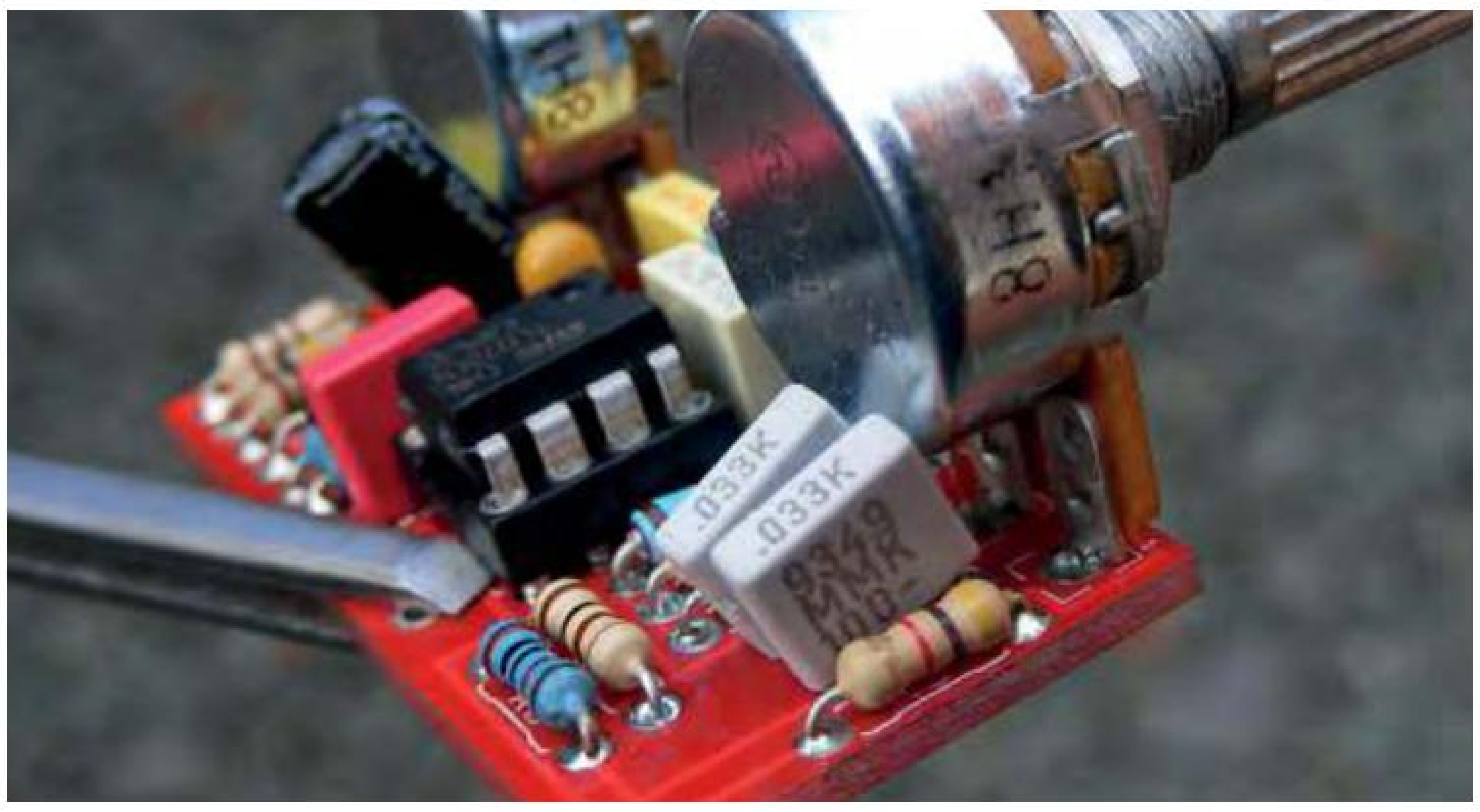


Fig.21. If you use standard vertical-mount pots then some capacitors may need to be bent over a bit. (Note the PCB files have now been stretched a bit to avoid this.)

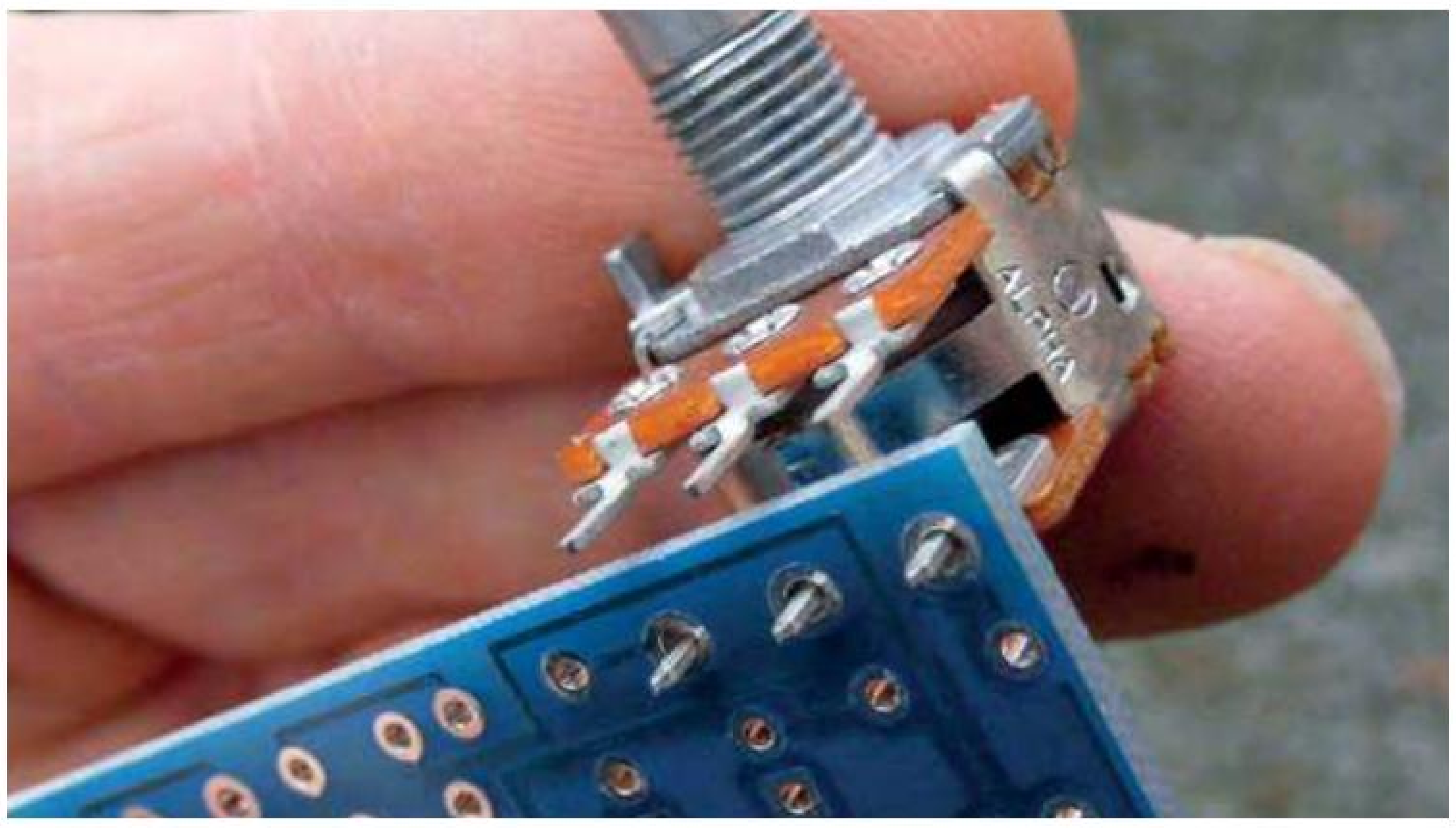


Fig.22. Dual-gang pots are needed for stereo and two boards are used with the pots only mounted on one. The other board can be linked with stiff wire.

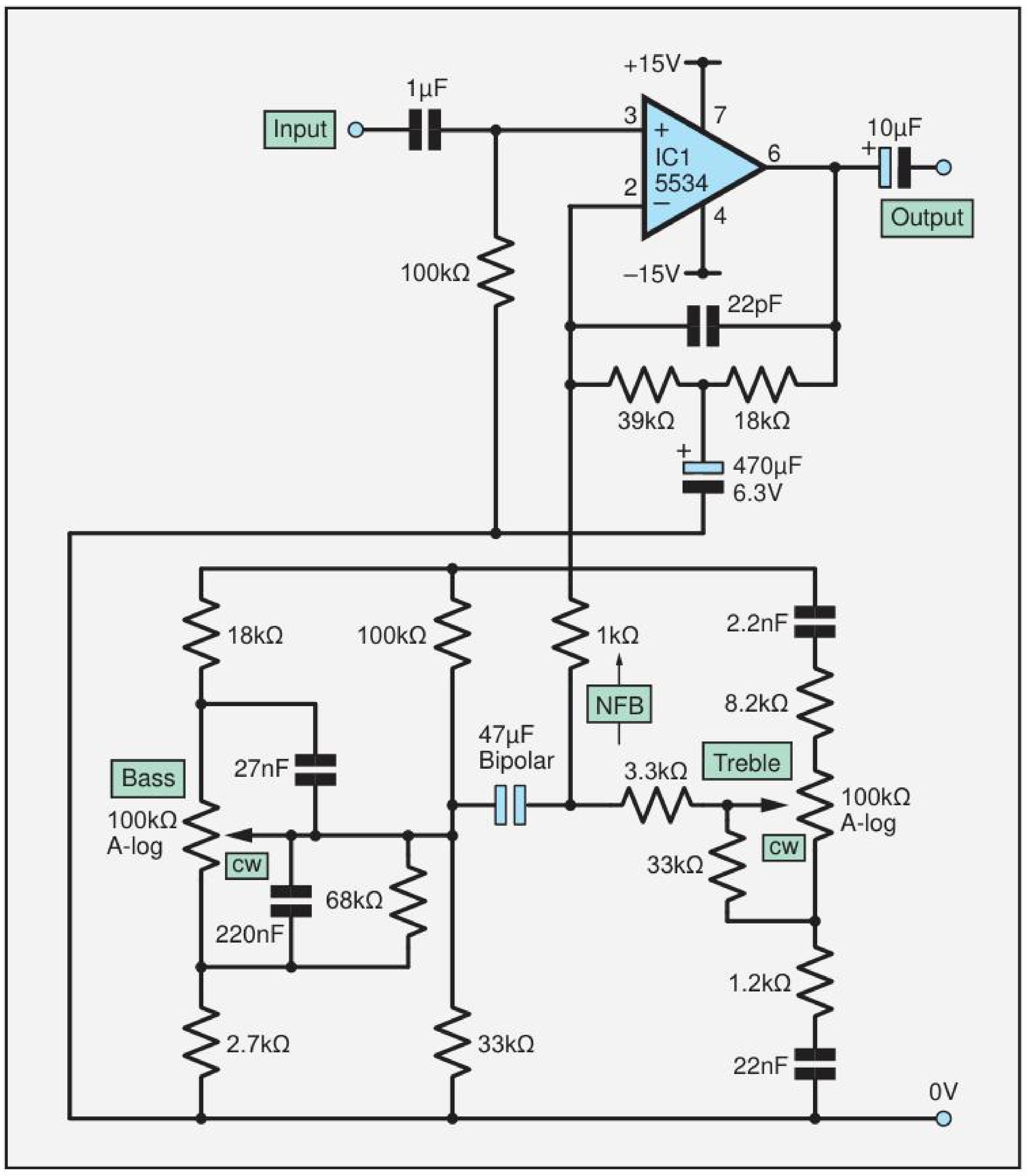
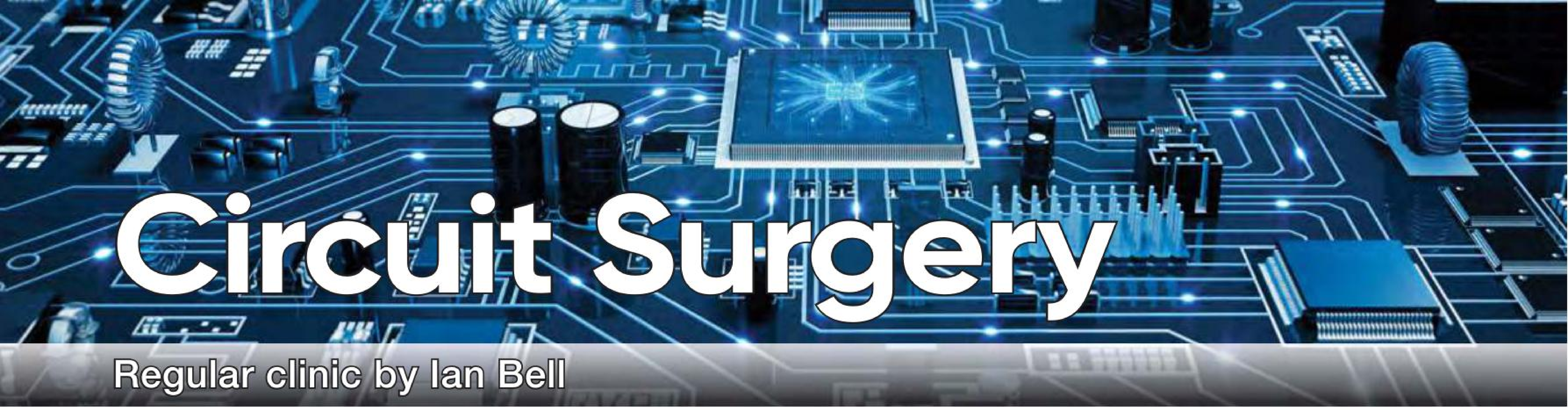


Fig.23. The 'Japanese' Baxandall tone control that is non-inverting, has high input impedance and thus needs no buffer. Only one op amp is needed, so good news if one wants to use a discrete amplifier circuit. Also, it has nice single pivot curves. On the down side, unfortunately, it uses unusual, (very) hard-to-find anti-log pots.



Circuit Surgery

Regular clinic by Ian Bell

Comparator circuit design

Recently, there have been a few posts about comparators on the EEWeb forum (**www.eeweb.com/forum/**). Qasim Ahtesham asked: 'Has anybody worked with an LM741 and used it for voltage comparison? I am trying to use one to compare two voltages, and use this as a thermostat. The issue is that once V+ > V– the output voltage is 1.3 to 1.6V, but not 0V. Do I need to add offset? I am only using the 5 pins at the moment. Any help on this, or a precise design for comparing the voltages will be highly appreciated.'

Graham Rounce posted: 'I'd like a simple way to determine if a voltage is outside of a certain band. Say, from +2V to +3V. I'd like a circuit with an output that goes high (5V) if the input voltage is =<2V or =>3V. Ideally, I'd like two separate outputs, one for =<2V and the other for =>3V. Op amp(s) will no doubt figure in the answer somewhere, but I get a bit confused when using them with a single (5V) supply.'

So, our next topic is comparators. We will look at their basic principles of operation, including the 'Schmitt trigger', or hysteresis, behaviour which is often set up in comparator circuits, and which has also been discussed in recent forum threads. Both initial posts mentioned op amps (the LM741 is an op amp), rather than comparator chips, although later in the discussion on Graham's question specific comparator ICs such as the LM393 are mentioned. Thus, one of the things we will look at is the difference between op amps, which can be used as comparators, and dedicated comparators. We will use LTspice simulations to illustrate circuit operation. In a recent series of *Circuit Surgery* articles (October 2018 to January 2019) we covered some basics of LTspice, but here we'll need some operations not covered there, such as simulating op amps that are not in the library. We will look at this separately next month, rather than explaining all aspects of the simulations here.

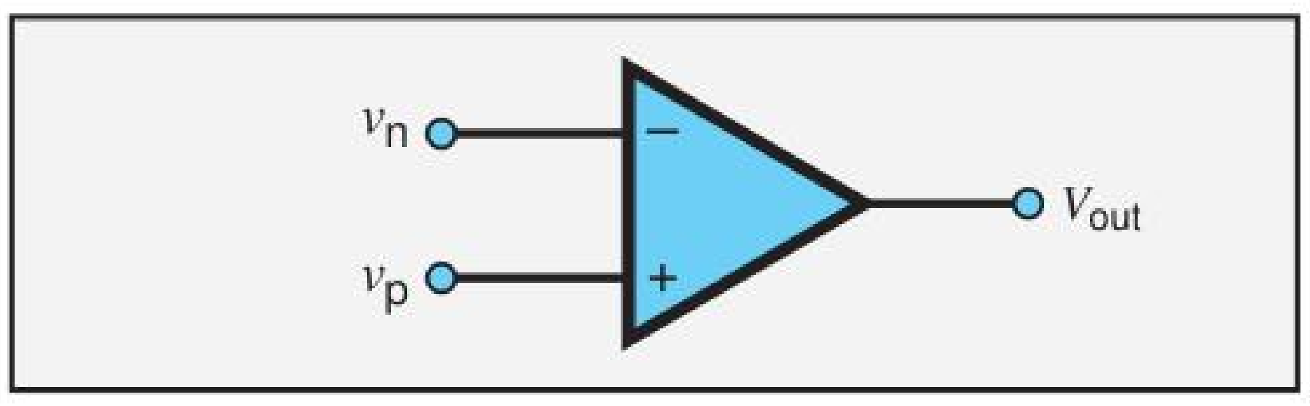


Fig.1. The symbol for a comparator – the same as for an op omp.

Definitions

A comparator is a circuit that compares one analogue signal with another and outputs a binary signal based on the result of the comparison. In effect, it is a one-bit analogue-to-digital converter. An op amp used without negative feedback (open loop) has very high gain, so for all but a small range of input voltage differences the output will be at the lowest or highest voltage available from the device (saturation), typically close to the supply rails. These two saturated output voltages may represent Boolean 0 and 1. An op amp used in this way behaves as a comparator. However, its performances may differ from that of a circuit specifically designed to act as a comparator. To compare the two it helps to first define comparator characteristics and see how these relate to op amps.

The circuit symbol for a comparator is shown in Fig.1 – at its most basic it is the same as an op amp's symbol. A comparator's output voltage may be written mathematically as follows:

$$V_{out} = \begin{cases} V_{OH} & \text{if} \quad v_p > v_n \quad \Rightarrow \text{logic } 1 \\ V_{OL} & \text{if} \quad v_p < v_n \quad \Rightarrow \text{logic } 0 \end{cases}$$

Where v_p and v_n are the input voltages, as shown on Fig.1 and V_{OH} is the logic 1, or high output voltage, and V_{OL} is the logic 0, or low output voltage.

The above equation implies infinite gain and zero offset, V_{OS}. This means that an infinitely small voltage change around $v_p = v_n$ will cause the output to switch (infinite gain), and that this switching will occur exactly at $v_p = v_n$ (zero offset).

Fig.2 shows the effect of finite gain and offset on a comparator's transfer characteristic (relationship between input voltage *difference* and output voltage). The effect of the offset and finite gain is to reduce the resolution of the comparator, so that the difference between the inputs must be larger than a certain minimum to give reliable detection. Fig.2 represents the situation for static inputs, where the resolution is approximately $V_{OS} + V_{IH} = V_{OS} + V_{OH}/gain$, but typically the resolution will get worse for changing input signals.

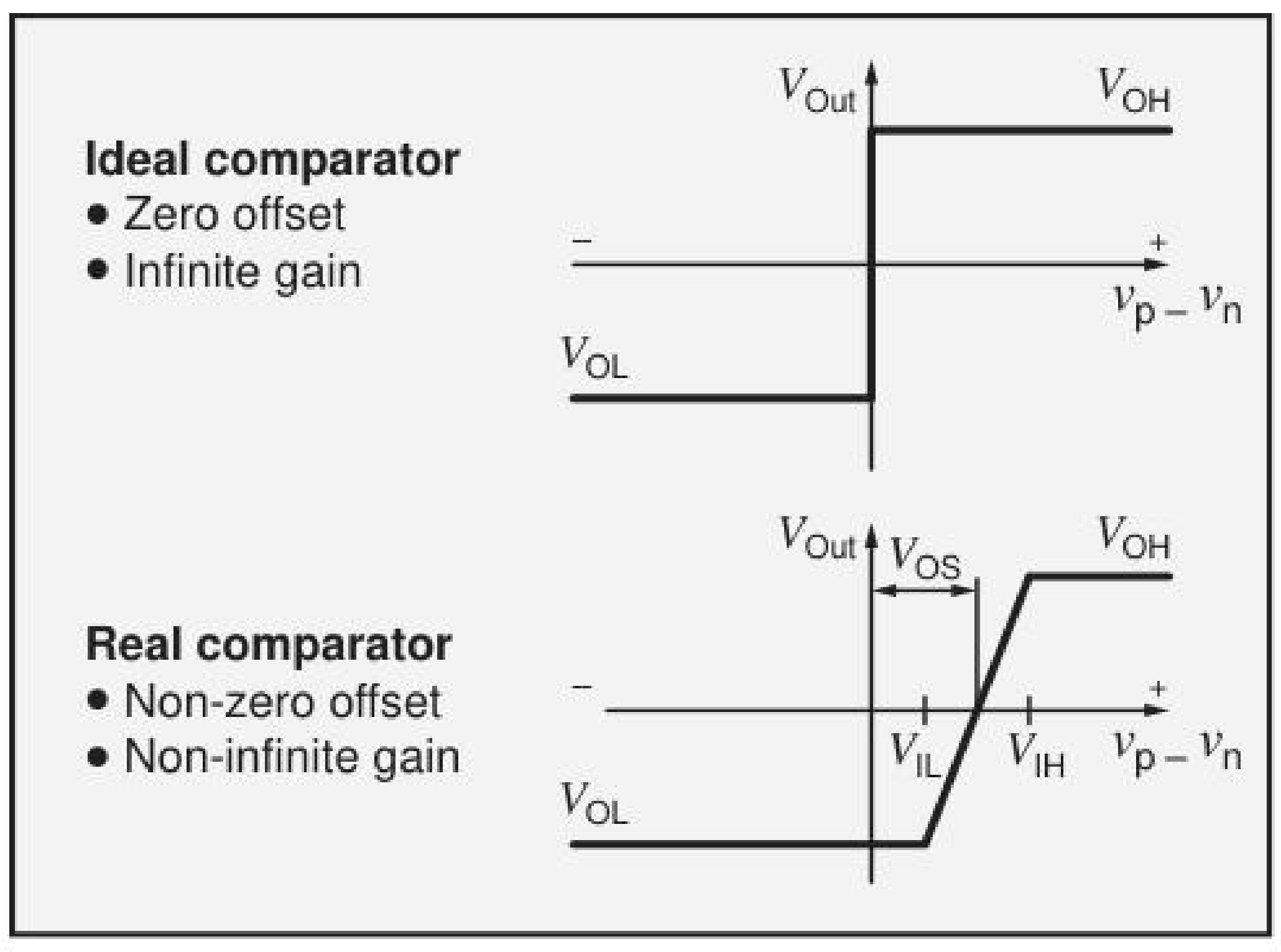


Fig.2. Comparator transfer characteristics.

Basic configurations

In many applications, comparators are used to compare an input signal with a reference voltage generated within the circuit. In such cases we can configure either inverting or non-inverting operation depending on which of the comparator's inputs is connected to the reference and the input (see Fig.3). A non-inverting comparator has a high output (logic 1) when the input is greater than the reference. For an inverting comparator the result is a high output when the input is below the reference.

Graham's requirement for a comparator circuit with two outputs, indicating if

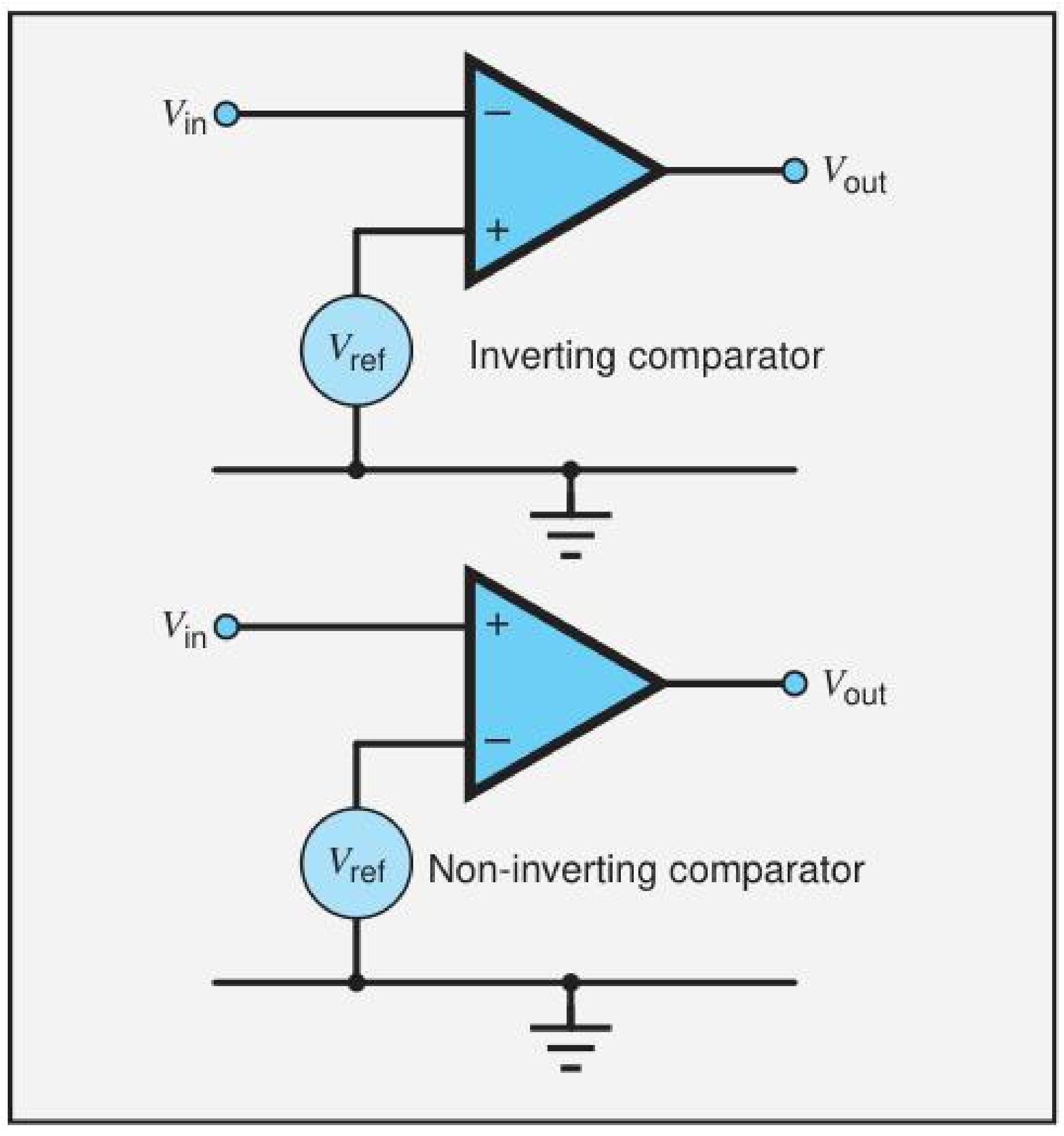


Fig.3. Inverting and non-inverting comparator configurations.

Simulation files

The LTSpice files discussed in *Circuit Surgery* are available for download from the *PE* website.

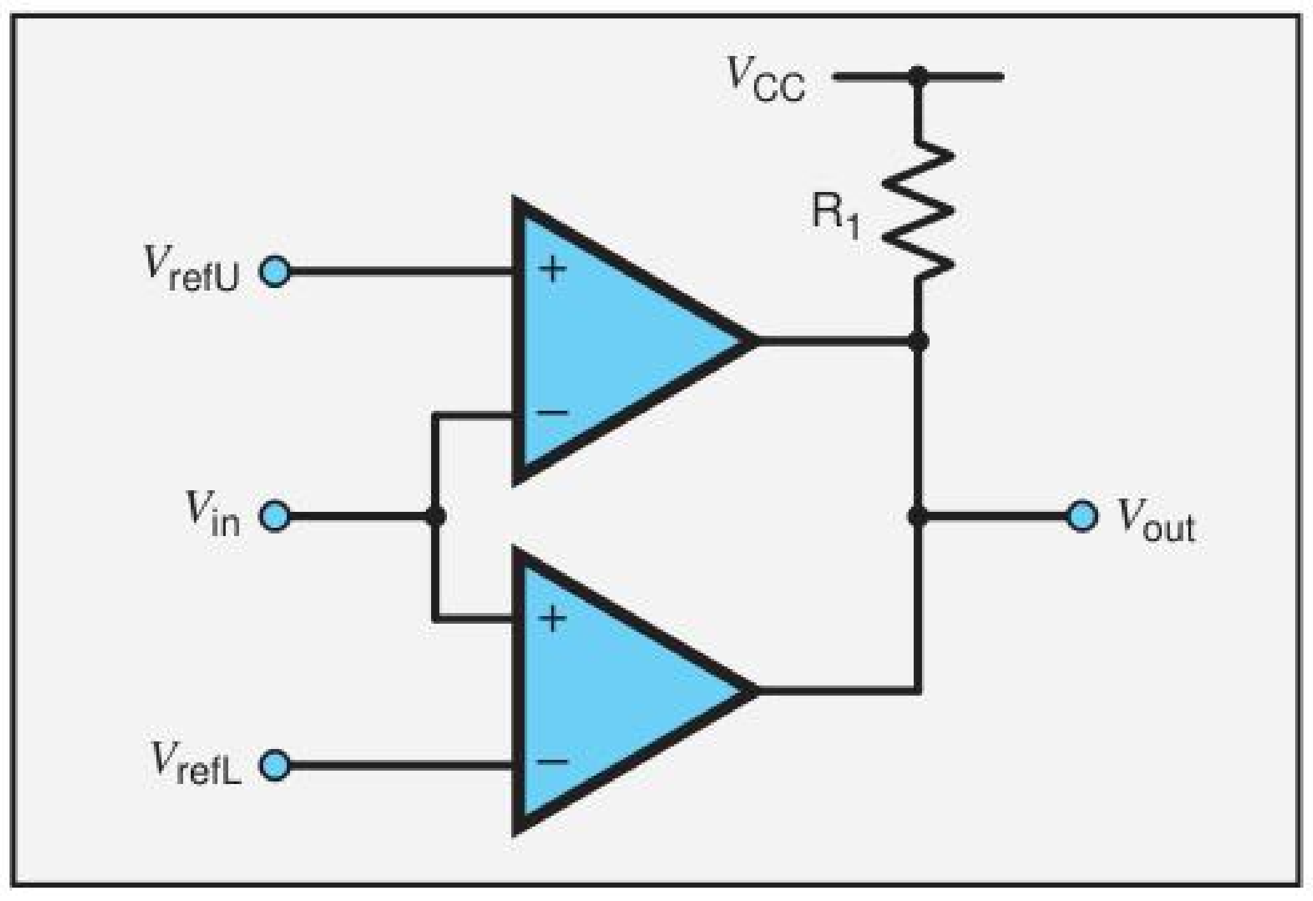


Fig.4. Window comparator using comparators with open-collector outputs.

the input voltage is ≤2V or ≥3V requires two comparators, one inverting (with a 2V reference) and the other non-inverting (with a 3V reference). A similar common requirement is to have a single output indicating when the input voltage is within a particular range (ie, above a lower limit V_{refL}, but below an upper limit V_{refU}). This is known as a 'window comparator'. It also requires two comparators (again, one inverting and one non-inverting), but the outputs are combined with a logical (Boolean) AND function (above-lower AND below-upper). The AND function can be implemented by a logic gate, but it is common to use comparators with open-collector outputs, which can be connected together to the same pull-up resistor (see Fig.4) to produce what is known as a 'wired-AND' circuit.

Op amps vs. comparators

Op amps are designed to be used with *negative* feedback (eg, provided by the resistors used to set the circuit gain). All amplifiers have some delay from input to output, which results in increasing phase shift as signal frequency increases. At some point, the phase shift reaches 180°, equivalent to inverting the signal, at which point the negative feedback network is actually delivering *positive* feedback. If the gain of the amplifier and feedback network together is greater than one at this frequency then oscillation will occur. The gain of most op amps is deliberately rolled off as frequency increases to prevent this instability – this is called compensation. Comparators are used open loop or with positive feedback, so compensation is not required, leading to significant differences between the two types of device.

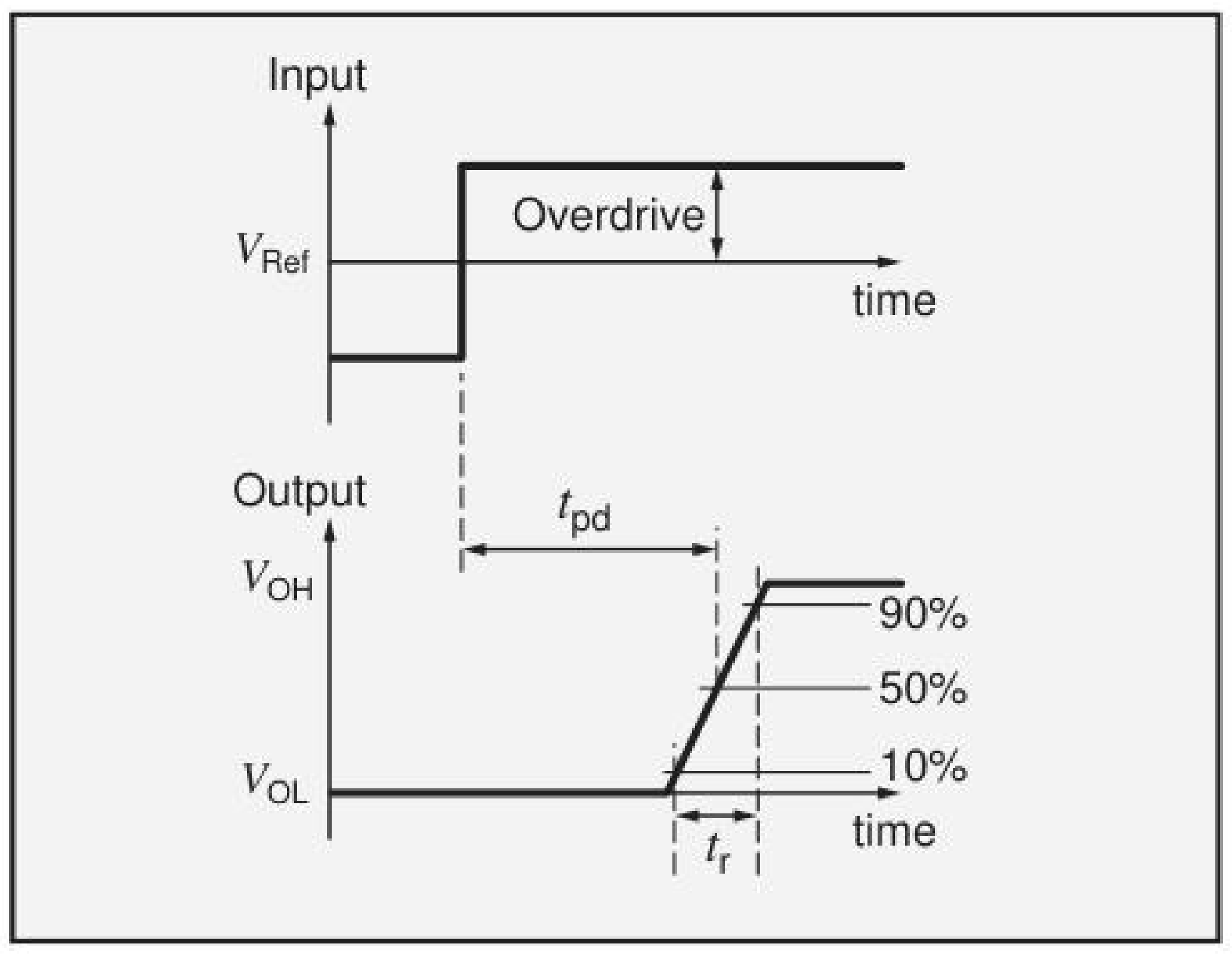


Fig.5. Comparator propagation delay.

Op amps are high-gain, linear, differential amplifiers; so in normal operation the voltage difference between an op amp's inputs is very small (typically microvolts to millivolts). Comparators often have much larger input differences. Not all op amps can tolerate large input voltage differences and they perform very poorly, or may even by damaged, under such conditions. Op amp input impedance may drop significantly for large input differences due to conduction of protection diodes – this could upset driving an op amp used as a comparator. Comparators are commonly used to compare voltages that are *not* close to half the supply range. For an op amp, this is a large common-mode input voltage. Again, not all op amps perform well under such conditions.

Gain and offset are characteristics shared by op amps and comparators; however, the switching behaviour of comparators means that they have characteristics related to switching which are not relevant to the standard analogue amplifier usage of op amps. The switching characteristics are illustrated in Fig.5, which shows comparator input and output waveforms for a non-inverting configuration with a fixed reference voltage.

Speed of switching

When the comparator input voltage crosses the reference voltage the comparator output will switch. This will not happen instantaneously – the time taken for the comparator output to reach 50% of the resulting voltage change is the propagation delay. The time taken for the comparator output voltage to rise from 10% to 90% of its range is the rise time. The amount of voltage applied to the comparator's input beyond the switching threshold (reference voltage) is known as the overdrive. Propagation delay and rise time are usually sensitive to overdrive, with increasing overdrive resulting in faster switching times. Comparator speed is also usually dependent on supply voltage.

The maximum rate of change of output voltage an op amp or comparator can deliver is the slew rate. Slew rate is important for op amps because it indicates how well the output voltage will track fast-changing analogue waveforms; failure to do so causes distortion. Slew rate also directly determines the maximum frequency at which an op amp can produce a pure sinewave at full output swing (the full-power bandwidth). However, sinewave output is of no relevance to comparators.

For any circuit used as a comparator, either the slew rate or the bandwidth may be the dominant factor in determining the propagation delay. Because comparators are just required to switch their outputs quickly, the slew rate itself is not usually very important as a specification – it is the propagation delay and rise time which are quoted. The compensation applied to op amps tends to reduce their slew rate, making them relatively slow when used as comparators.

A comparator's output will typically switch between the positive and negative supply voltages (or ground and supply in single-supply circuits). However, the output may switch, or it may be possible to arrange for it to switch, to a different voltage from the main comparator supply to facilitate interfacing to logic circuits. Often, comparator output circuits are designed to be easy to interface with specific types of logic. Comparators are therefore available with a variety of output configurations including push-pull, open-drain or collector and LVDS (low-voltage differential signalling). Open-drain and open-collector outputs require an external resistor connected from the output to the positive (digital) supply.

Op amps are designed for use where the output voltage does not hit the supply rails – this would normally imply clipping of the waveform and hence distortion. When op amp outputs are driven hard into saturation they tend to be slow to recover. Like compensation, this makes op amps poor comparators where fast switching is required. The internal circuitry of comparators is designed to prevent the output stages going far into saturation, allowing them to recover very quickly. A further subtlety to this is that op amp saturation recovery time is likely to vary between individual devices, making the propagation delay somewhat unpredictable.

LM393 vs. LM741 simulation

Qasim asked about using the LM741 as a comparator. If we obtain a SPICE models for this op amp and a comparator then we can run a simulation to compare output waveforms. Fig.6 shows an LTspice schematic for just this purpose (we will discuss the setup of this next month). The simulation configures both devices in a basic comparator configuration on a single supply, which is 20V to correspond with the LM741 recommended minimum of ±10V. The lack of a split supply is not an issue with these basic circuits and a comparator reference at the centre of the supply range is used. The LM393 has an open-collector output so a pull-up resistor (R1) is required.

The results of the simulation are shown in Fig.7. This shows that the output of the LM741 only gets to within about ±1V of the supplies – as described by Qasim. This is an inherent characteristic of the LM741, but not fundamental to using op amps as comparators – rail-to-rail output op amps are available, which can output voltages closer to the supply rails. The LM393's output goes up to 20V because there is no load, and down to 0.15V (the collector-emitter saturation voltage of the

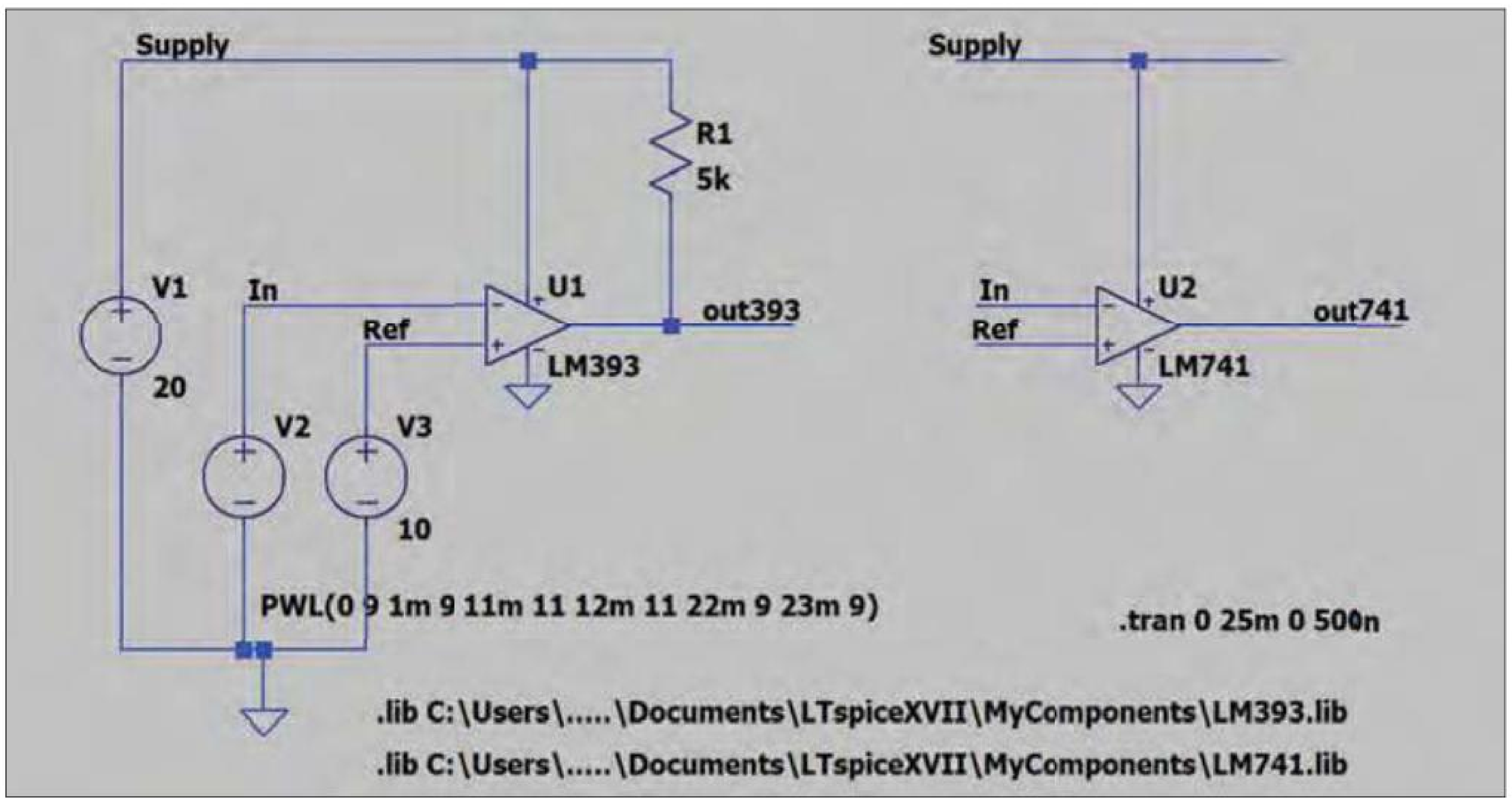

Supply
Supply
R1
5k
V1
20
In
Ref
U1
LM393
out393
V2
V3
10
In
Ref
U2
LM741
out741
PWL(0 9 1m 9 11m 11 12m 11 22m 9 23m 9)
.tran 0 25m 0 500n
.lib C:\Users\.....\Documents\LTspiceXVII\MyComponents\LM393.lib
.lib C:\Users\.....\Documents\LTspiceXVII\MyComponents\LM741.lib

Fig.6. LTspice simulator schematic for comparing an LM393 comparator with an LM741 op amp used as a comparator. This simulation's setup will be discussed next month.

open-collector output transistor). The simulation also shows the LM741 op amp responds much more slowly than the LM393 comparator, as discussed above.

In summary, op amps can be used as comparators, but not without difficulties, and only in relatively slow-speed applications. Op amps may not perform well with large differential or common-mode inputs, but such conditions are common in comparator applications – so the op amp's datasheet should be consulted for relevant capabilities. Interfacing an op amp output to logic may also be less straightforward than with a comparator specifically designed to drive logic inputs.

Hysteresis

A comparator used with a single threshold (reference) value, as in the preceding examples, may switch states many times as a noisy, slowly changing input crosses the threshold. This is often undesirable; for example, if the number of threshold-crossings is to be counted or you want to avoid 'chattering' when the input is close to the threshold. The problem may be overcome by using two thresholds; eg, V_{TH} and V_{TL}. The difference between V_{TH} and V_{TL} is called the 'hysteresis'. A comparator with hysteresis can be made by applying positive feedback to a basic comparator to shift the threshold slightly, depending on which of its two output states the comparator is currently in. Comparators with hysteresis are also known as 'regenerative comparators' or 'Schmitt triggers'.

The input-output response of a comparator with hysteresis is shown in Fig.8. When the input *increases* past V_{TH} the comparator switches, but it does not switch back if the input decreases back past V_{TH}. The input must decrease past V_{TL} before the comparator switches again.

If the input noise level is known, the hysteresis can be set slightly larger than this. The comparator will then not switch as a result of the noise. Fig.9 shows an LTspice simulation schematic for comparing comparator circuits with and without hysteresis. The comparators use generic op amp models (UniversalOpamps2), which are provided with LTspice, rather than specific devices like the LM393 or LM741. This is sufficient to produce waveforms to illustrate a general situation. LTspice techniques used here will be discussed next month.

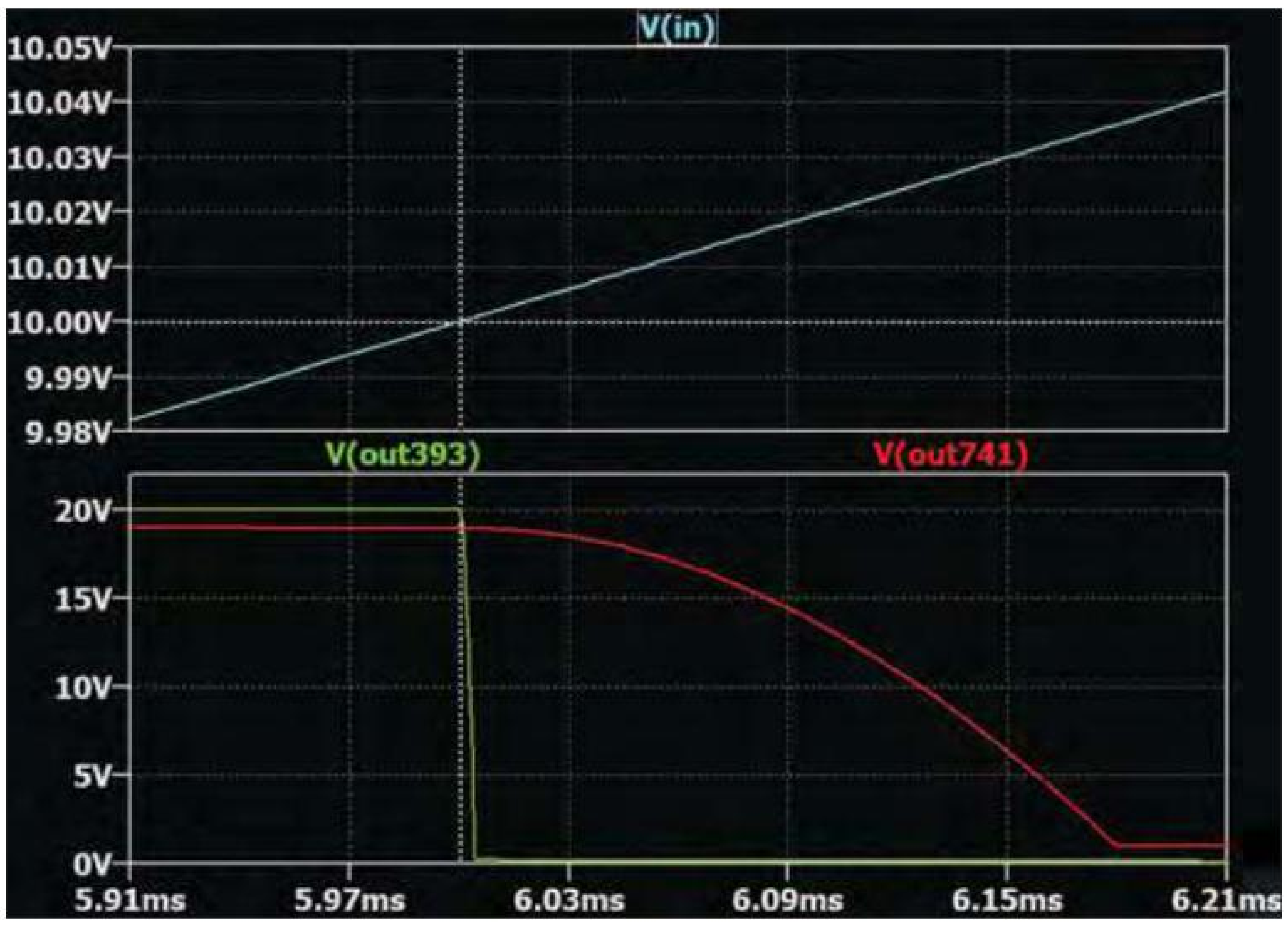


Fig.7. Results from simulation of the circuit shown in Fig.6. The cursor (white dotted line) marks where the threshold is crossed.

Fig.10 and Fig.11 show the result of applying the same noisy signal to a basic comparator (the circuit using U2 in Fig.9) and one with hysteresis (the U1 circuit). As mentioned above, the basic comparator switches multiple times as the noisy signal crosses the threshold, whereas the comparator with hysteresis switches cleanly. Fig.10 shows that the behaviour of the basic comparator as it switches is different each time due to the random nature of the noise. Fig.11 zooms into one threshold crossing.

Circuit design

As shown in Fig.12, a comparator with hysteresis can be made using a basic comparator with positive feedback. As the thresholds depend on the comparator's output voltages these should ideally be accurately controlled.

Refer to Fig.12. The switching point V_{comp} depends on V_{ref} and V_{out}. V_{ref} will usually be fixed but V_{out} depends on the current state of the comparator. V_{out} can take one of two values, which we will assume to be $\pm V_O$ (the positive and negative outputs apply to split-supply circuits). Initially, let us assume that $V_{in} < V_{comp}$ so $V_{out} = +V_O$ (note that V_{in} goes to the inverting input). If V_{in} is slowly increased this condition remains until $V_{in} = V_{comp} = V_{TH}$ (see Fig.7), where:

$$V_{TH} = \frac{R_2}{R_1 + R_2}V_{ref} + \frac{R_1}{R_1 + R_2}V_O$$

This equation is obtained by applying the potential divider equation twice and adding the results. First to find the contribution of V_{ref} to V_{comp} with $V_{out} = 0$ and then to find the contribution of V_{out} to V_{comp} with $V_{ref} = 0$. This is an application of the superposition theorem.

On switching at $V_{comp} = V_A$ the output changes to $V_{out} = -V_O$, changing the switching point to a new value, $V_{comp} = V_B$

$$V_{TL} = \frac{R_2}{R_1 + R_2}V_{ref} - \frac{R_1}{R_1 + R_2}V_O$$

V_{out} will now stay at $-V_O$ until the input falls below V_{comp} again. The difference in the switching points; ie, the hysteresis (V_H) is given by:

$$V_H = V_{TH} - V_{TL} = \frac{2R_1}{R_1 + R_2}V_O$$

The switching is symmetrical about the average of V_A and V_B, which is:

$$\frac{R_2}{R_1 + R_2}V_{ref}$$

If R_2 is much larger than R_1 (which is common) the average of V_A and V_B is approximately equal to V_{ref}. Under these conditions the comparator switches at points $V_H/2$ above and below V_{ref}. For the circuit in Fig.9 the switching points are 1.548 ±0.161V (1.71V and 1.39V). Note, this

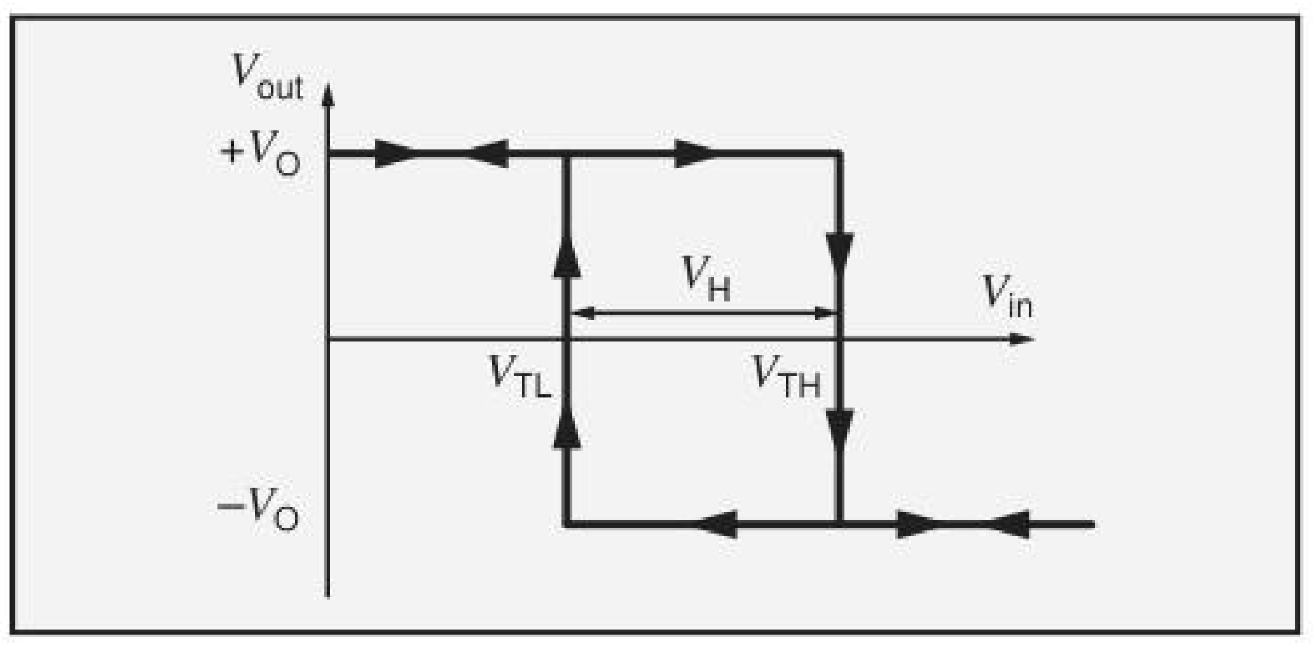


Fig.8. Switching characteristic of comparator with hysteresis.

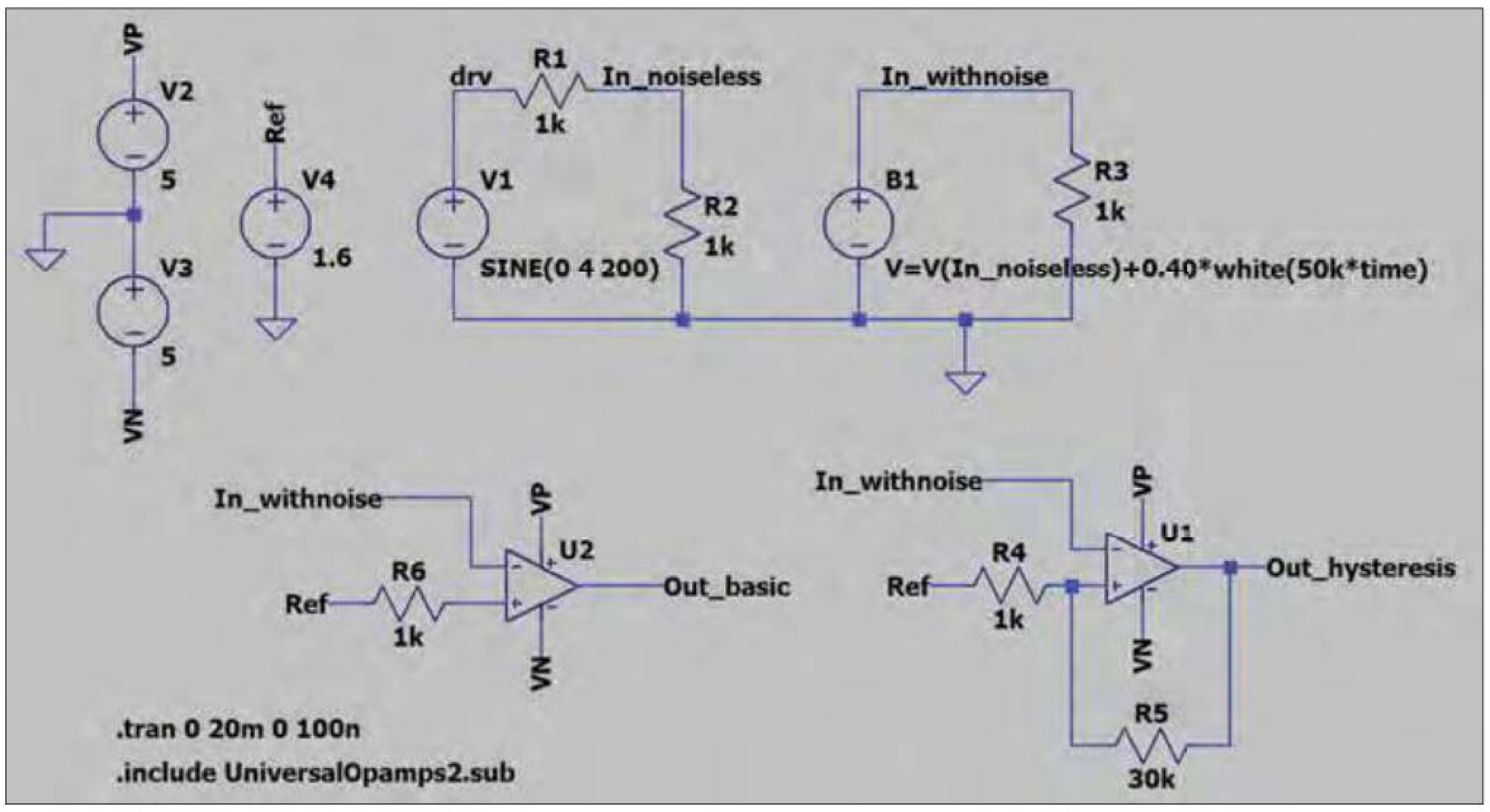


Fig.9. LTspice simulation schematic using a simple generic op amp model as a comparator to show the difference between circuits with and without hysteresis. The setup of this simulation will be discussed next month.

is not centred exactly on the 1.6V reference.

If the comparator is run from a single supply the output (ideally) switches between $+V_O$ and 0V rather than $+V_O$ and $-V_O$. The formula for V_{TH} is the same, but the zero removes the $-V_O$ term from the V_{TL} equation, so it becomes:

$$V_{TL} = \frac{R_2}{R_1 + R_2} V_{ref}$$

If R_2 is much larger than R_1 the lower threshold (V_{TL}) is close to V_{ref}, rather than V_{ref} being approximately the centre between the two thresholds, as it is with the split supply circuit. This may be a source of confusion with single-supply versions of the circuit.

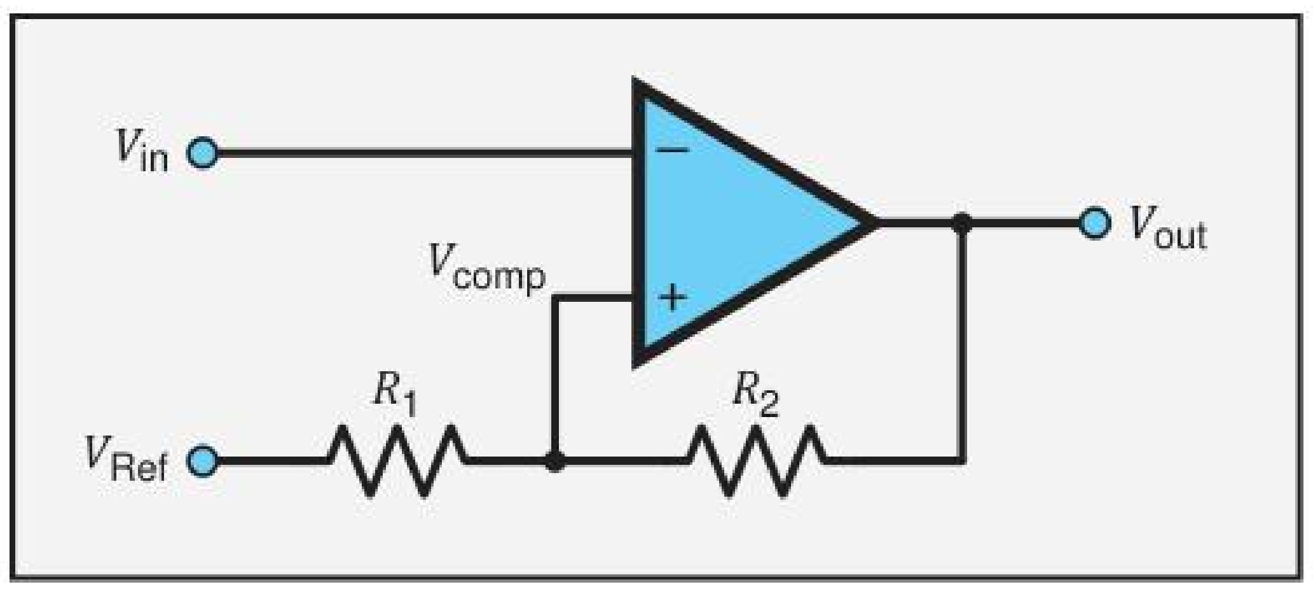


Fig.12. Using positive feedback to apply hysteresis to a comparator.

Potential dividers

V_{ref} in Fig.12 can be obtained from a potential divider connected between the supplies. Ideally, in order to prevent the threshold setting network (R_1 and R_2 in Fig.12) from loading the potential divider and influencing the reference voltage, the potential resistors should have relatively low resistance. Alternatively, interaction between a potential divider and feedback resistor can be used to set the two threshold voltages using the circuit shown in Fig.13.

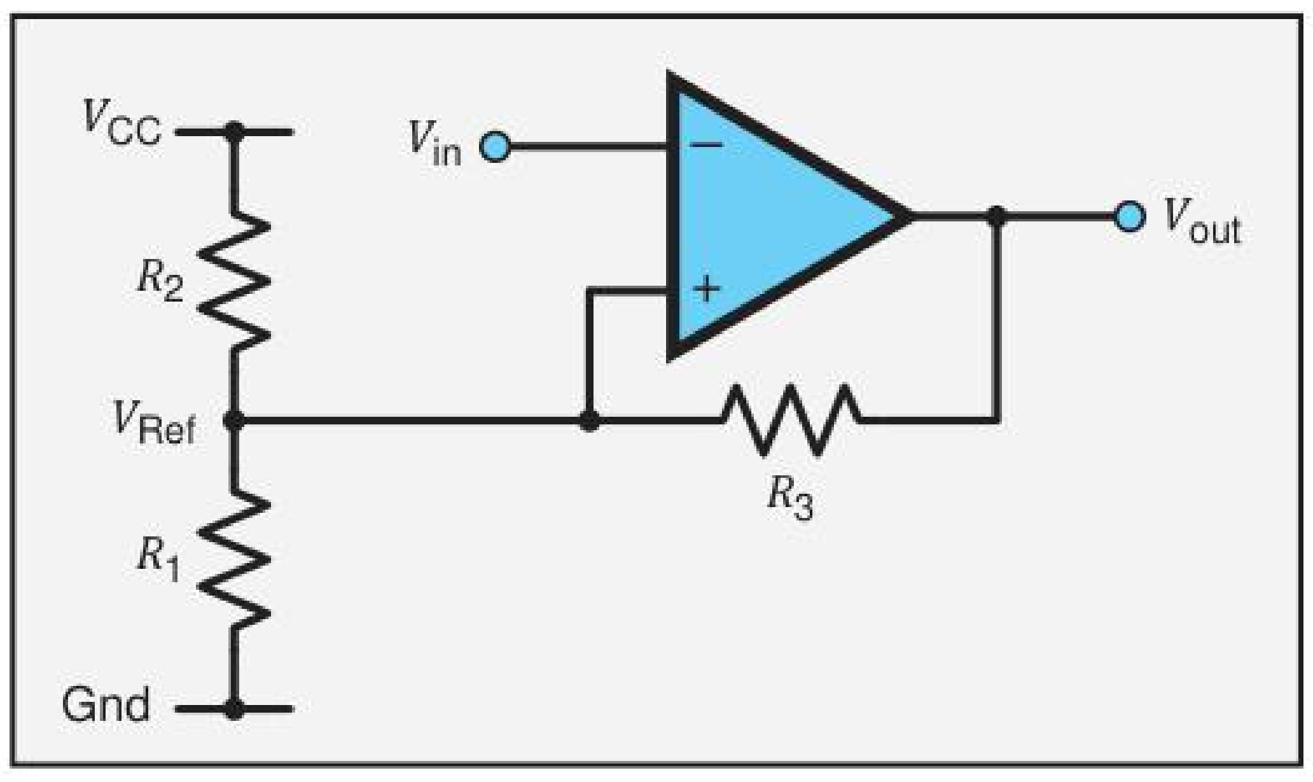


Fig.13. A potential divider can be used in conjunction with a feedback resistor to set comparator thresholds. (Single supply circuit shown.)

If we run the circuit in Fig.13 from a single supply, and assume the comparator's output voltage is 'rail to rail'($+V_{CC}$ or 0V), then it is reasonably straightforward to find the two thresholds. When the comparator output is 0V, R_3 is effectively connected to ground in parallel with R_1. The parallel value of these resistors, R_{P13}, is, using the formula for two parallel resistors:

$$R_{P13} = \frac{R_1 R_3}{R_1 + R_3}$$

This resistance forms a potential divider with R_2 to set the lower threshold:

$$V_{TL} = \frac{R_{P13}}{R_{P13} + R_2} V_{CC}$$

When the comparator output is V_{CC}, R_3 is effectively connected to the supply in parallel with R_2. The parallel value of these resistors, R_{P23}, is given by:

$$R_{P23} = \frac{R_2 R_3}{R_2 + R_3}$$

This resistance forms a potential divider with R_1 to set the upper threshold:

$$V_{TH} = \frac{R_1}{R_1 + R_{P23}} V_{CC}$$

If the output is not perfectly rail to rail then we need more complex equations which include the two comparator output voltages as well as V_{CC}.

The switching points for the circuit in Fig.13 are not necessarily symmetrical around the open circuit voltage of the $R1$-$R2$ potential divider. For example, if we have V_{CC} = 5V and choose R_1 = 5kΩ and R_2 = 25kΩ we get a potential divider voltage of 833mV, with nothing else connected to it. If we have a feedback resistor R_3 = 100kΩ then R_{P13} is 4.762kΩ, R_{P23} is 20kΩ and the threshold voltages are V_{TL} = 0.800V, and V_{TH} = 1.00V (using the above formula). We have a total hysteresis of 200mV, but the lower threshold (in this example) is much closer to the open-circuit potential divider voltage than the upper threshold.

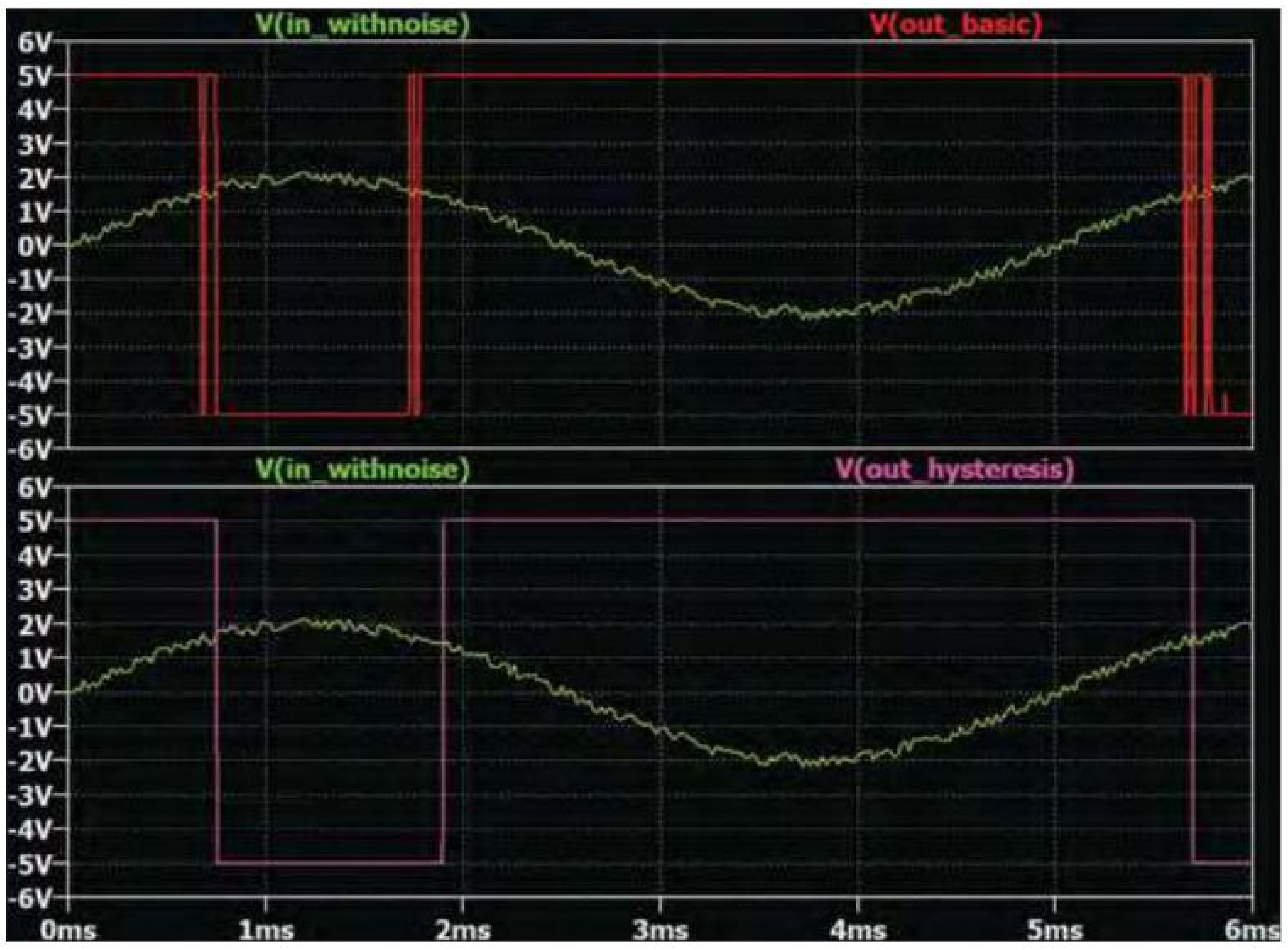


Fig.10. Response of the two comparators from simulation schematic in Fig.8.

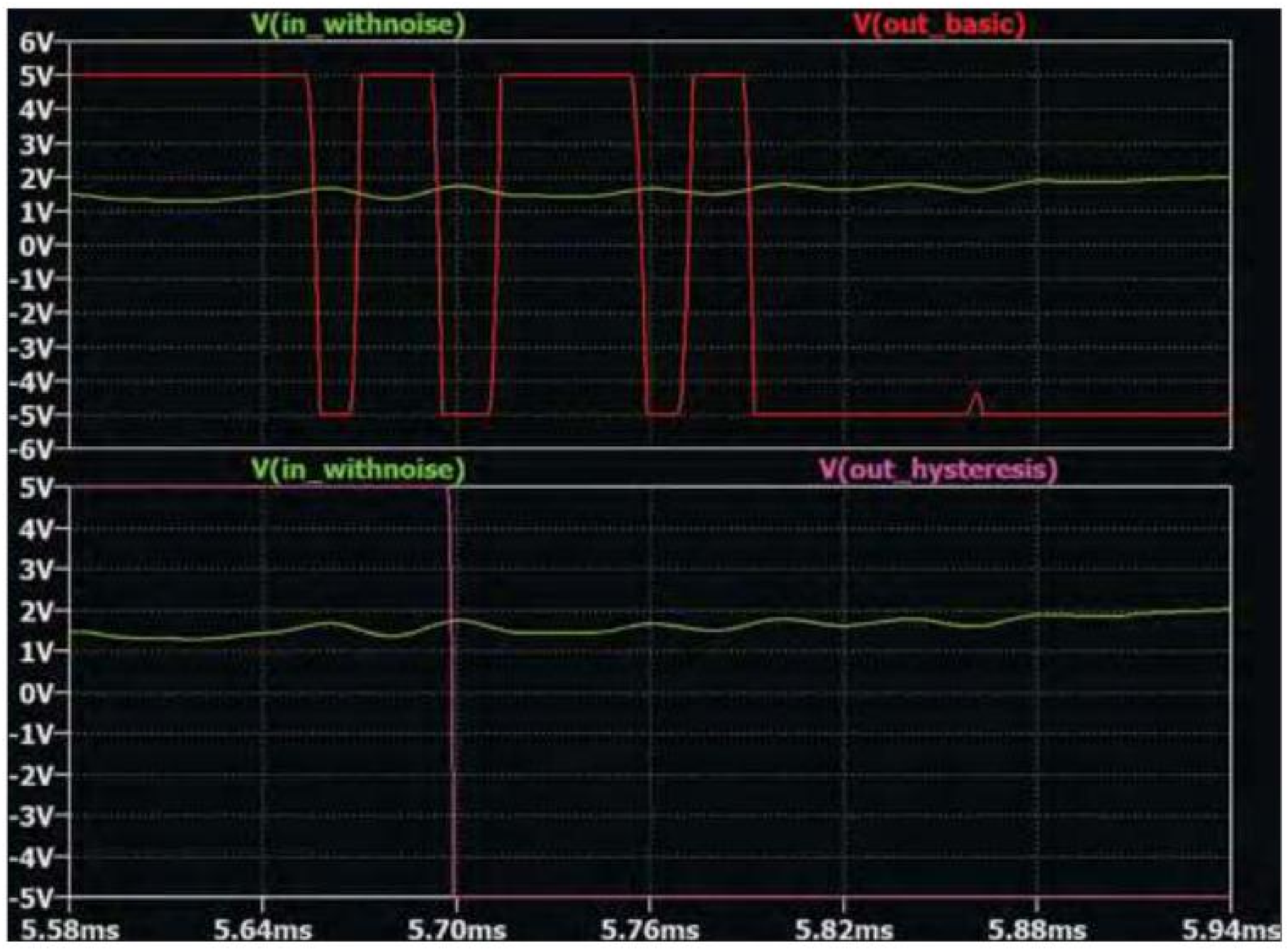


Fig.11. A zoomed in view of one of the threshold crossings from Fig.10 to show the waveform at the switching point in more detail.

PICn' Mix

Mike Hibbett's column for PIC project enlightenment and related topics

Small, cheap and powerful – Part 1

This month, we go back to basics, looking at Microchip's smallest and most feature-limited processor, the PIC10F200 family. This is a tiny processor available in a hobbyist-friendly eight-pin DIL package or even a six-pin SMD (surface mount device) part not much bigger than a speck of dust, as can been seen in Fig.1. In this series of articles, we will use both types of packages. Example projects will be created on a range of board types, from breadboard and Veroboard to tiny PCBs using the SMD version.

Minimalist PIC features

So, what does this device consist of? It's offered in a range of packages, but we will focus on the two hobbyist friendly parts – the eight-pin PDIP (plastic dual in line package) and the six-pin surface mount, shown in Fig.1. Both packages offer the same functionality – on the eight-pin variant, two pins are not connected.

These devices can operate between 2.0V and 5.5V, which makes them ideal for use with two or three alkaline cells without the additional cost of a voltage regulator. Current consumption is 250μA at 2.0V, up to 1mA at 5.5V. The processor runs at a fixed frequency of 4MHz derived from an internal RC (resistor-capacitor) oscillator. An RC oscillator is normally very inaccurate, but each device is calibrated during manufacture and a calibration correction value is programmed into a word in Flash memory. It's amazing that they can afford the cost of bespoke timing correction given the low cost of the chip – such is the power and efficiency of the automation machines used during chip manufacturing.

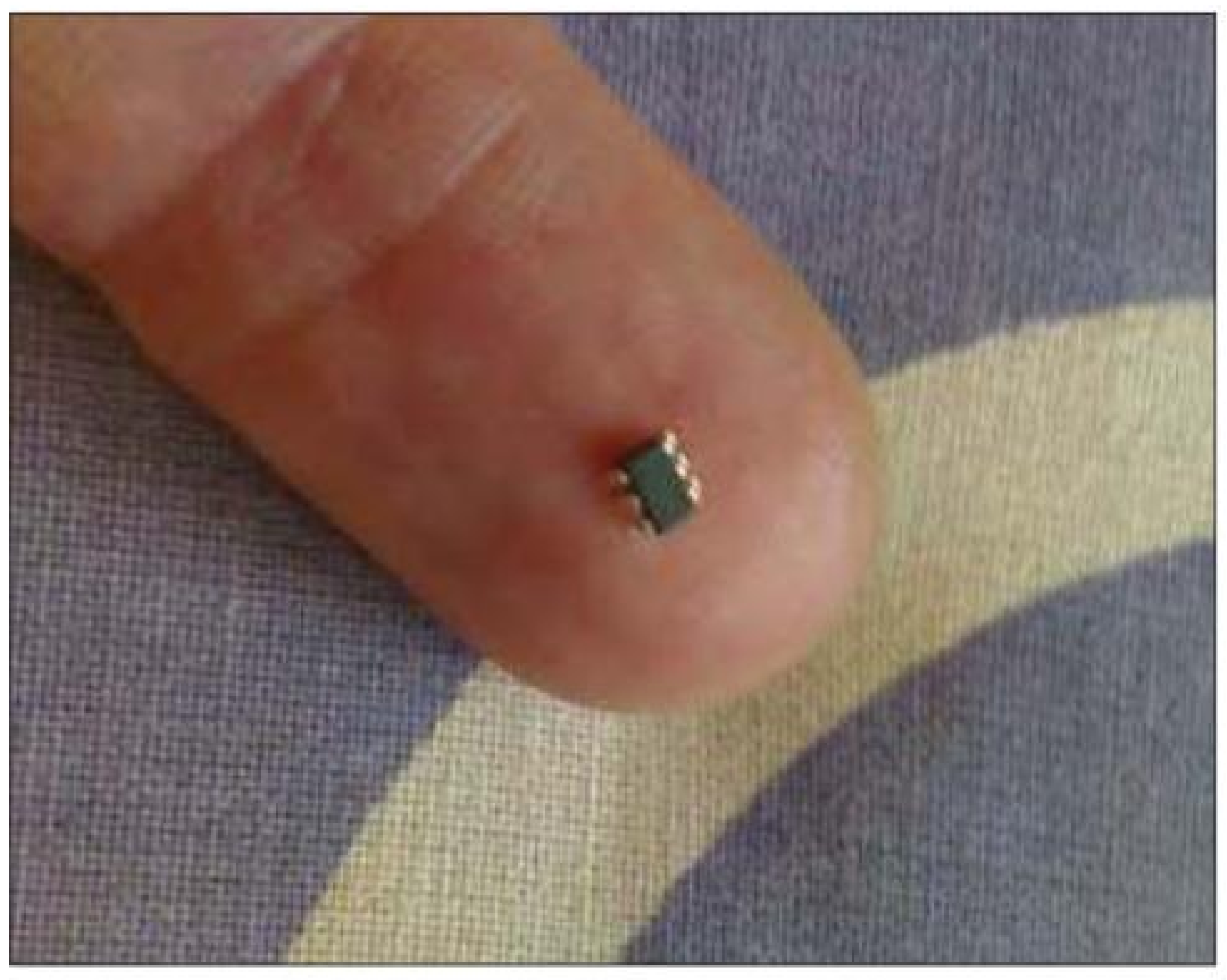


Fig.1. The PIC six-pin SMD package – 'smart dust'!

The device has an active standby mode – sleep – where it can remain inactive with it's processor clock turned off, but can monitor for transitions on GPIO input pins and wake immediately. In this mode the processor consumes just 200nA – a figure that is as close to nothing as you can get!

The processor is offered in four variants, or 'SKU' varieties, the standard industry term. The features included in each variant are shown in Table 1. Of the four variants we will focus on the PIC10F202; the most limited feature set, but with a little more memory to give us some flexibility. To summarise, the capabilities of the processor we are playing with are:

- 512 words of code, 24 bytes of RAM
- Internal oscillator calibrated to 1%
- 8-bit timer
- 4 I/O pins
- Wake on pin change
- Very low power consumption

'Wait', I hear you say. 'Is that all? What about I^{2}C bus, SPI bus, UART?' All of these functions can be implemented using 'bit bashing' of pins. If the protocol that runs above those interface requirements is not too demanding, then there is a chance the PIC10F will be capable of achieving them. Remember, if your needs are greater, simply choose a more powerful processor. I once asked a Microchip account manager how many processor SKUs they had. He went away for a bit and then came back saying, 'I gave up at 17,000'. The PIC10F family simply represents the absolute lowest-cost options available from them, and we are going to push those limits – hard!

I^{2}C and SPI bus communications are serial protocols based on 'handshakes' between devices to pass data. They do not rely on accurate clock signals or accurate timing of each data bit. UART communication, however, does. With the calibration data provided by Microchip during each device's manufacture, the device's clock can be set to an accuracy of 1% – enough for UART communication at reasonable speeds. Given this addition, proved by Microchip, the devices capabilities can be considered to be:

Table 1. Feature sets of the PIC10F200 family

Feature	PIC10F200	PIC10F202	PIC10F204	PIC10F206
Maximum frequency of operation	4	4	4	4
Flash program memory	256	512	256	512
Data memory (bytes)	16	24	16	24
Timer module(s)	TMR0	TMR0	TMR0	TMR0
Wake-up from sleep on pin change	Yes	Yes	Yes	Yes
Comparators	0	0	1	1
I/O pins	3	3	3	3
Input-only pins	1	1	1	1
Internal pull-ups	Yes	Yes	Yes	Yes
In-circuit serial programing	Yes	Yes	Yes	Yes
Number of instructions	33	33	33	33

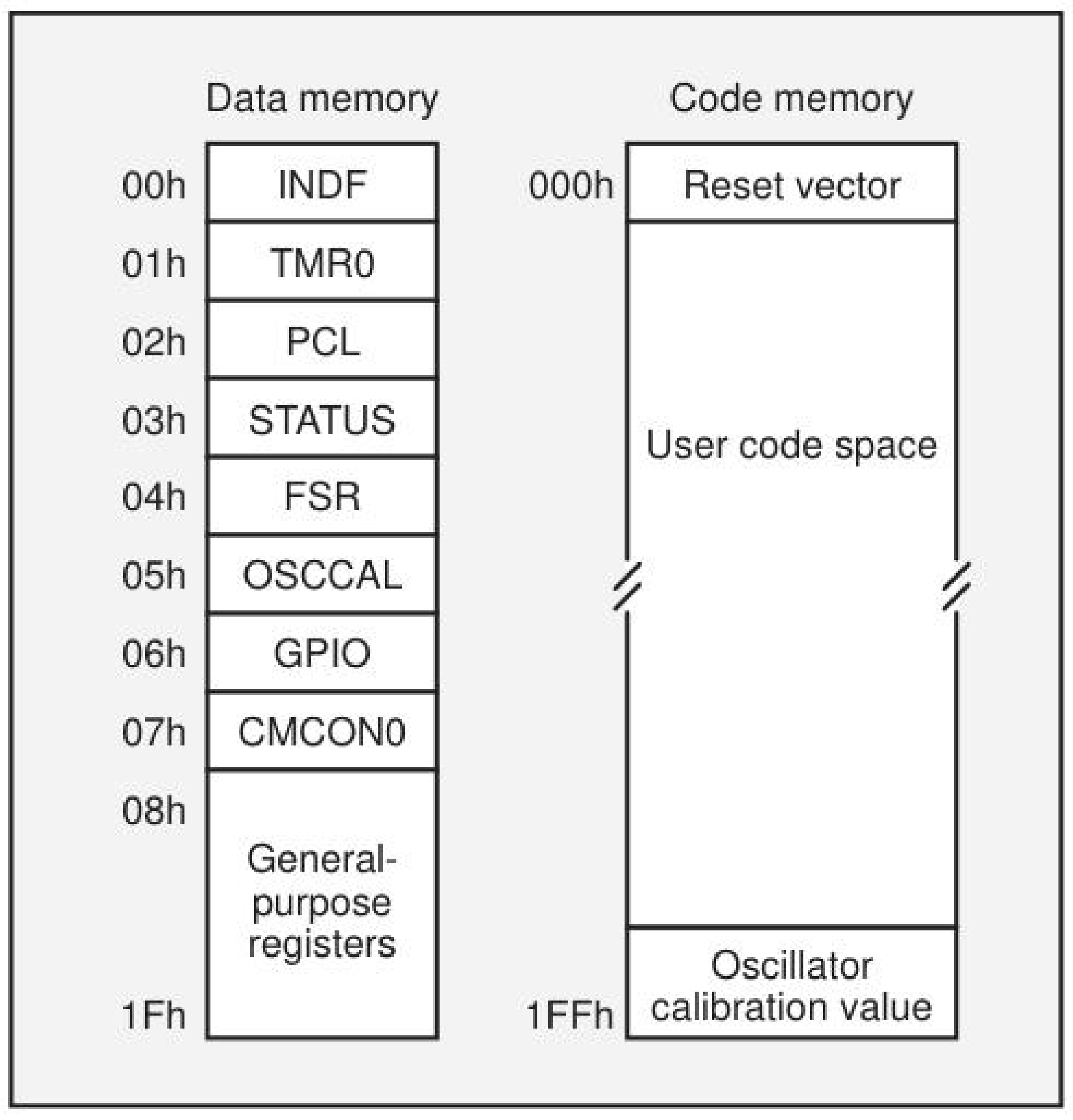


Fig.2. PIC10F202 memory layout

- 512 words of code, 24 bytes of RAM
- An internal oscillator calibrated to 1%
- 8-bit timer
- 4 I/O pins
- Wake on pin change
- SPI interface
- I^{2}C interface
- UART interface
- 1-wire interface
- Very low power consumption.

This now makes the processor a very capable device – so long as your actual requirements on what the device does with those capabilities are not too demanding. Needless to say, you are not going to be able to drive an LCD display with a parallel interface. But you could drive a display equipped with a serial interface, so long as you do not need to display lots of text. All of that text has to be stored in memory, and we have just 512 bytes for Flash memory storage. A serial LCD showing the time, or perhaps the ambient temperature, is probably feasible. Offering both? Perhaps not – but half the fun is experimenting and squeezing the maximum capability from limited resources!

The memory layout of the device is shown in Fig.2. Along with the general-purpose registers – our available RAM for program variables – there are just eight special function registers. Here lies the PIC10F's attraction for other reasons – it is a very simple device, and therefore makes an ideal basis for learning the complexities of a microcontroller. Other systems, even the venerable Arduino processor, contain far more registers, which may overwhelm a student when trying to master the lower levels of microcontroller systems. The PIC10F makes an ideal entry point.

The pin-out of the first device we will be using is shown in Fig.3. This variant is a dual-in-line plastic DIP package, which is easy to solder to Veroboard, or plug into a breadboard. The pin-out shows how several functions are mapped to the I/O pins.

Possibilities

So, what can you actually do with a processor that is so limited, and with so little memory?

First, remember that it is a very small device – as can be seen in Fig.1. This means it can be retrofitted into an existing device or be used to create a tiny functional product.

So what kind of projects are possible? Here are some examples that spring to mind that we believe are possible.

Smart bike light

Fitted with an accelerometer IC, the PIC10F could be used to sense sudden deceleration, and flash the brake light automatically. A safety feature perhaps?

Disco lights

A string of RGB LEDs that change colour depending on motion detected by an acceleration IC. Could be fun on the dance floor when mounted on your best dancing shoes!

Logic probe

An improvement on an old classic – a probe that can indicate when high, low or pulses are detected on a wire. Combined with an RGB LED for a more modern approach to indicating the state. We could program green for low, red for high and blue for pulse.

Temperature warning

A small device that can be placed in the refrigerator and warn when the temperature rises too high. With some additional components it should be possible to detect a low-battery condition too, and provide an additional warning. A circuit like this could run from a coin cell battery and be realised in a very small enclosure.

Knock-knock switch

Making use of a simple piezo buzzer to act as a vibration sensor, create a device that will toggle a relay when a 'knock-knock' physical signal is detected on a surface. Ideal for turning desk lights on and off, or as a general novelty switch.

Smart switch

A relay-controlled power source that is controlled by a push switch. The switch is de-bounced for a period of time before turning on. While on, a long press of the button will flash an LED as a warning, and then power the relay off if continued to be held. If a latching relay is used, this circuit will consume virtually no power, making it an ideal way of controlling battery-powered devices. It could also have a timer for 'auto power-off' functionality.

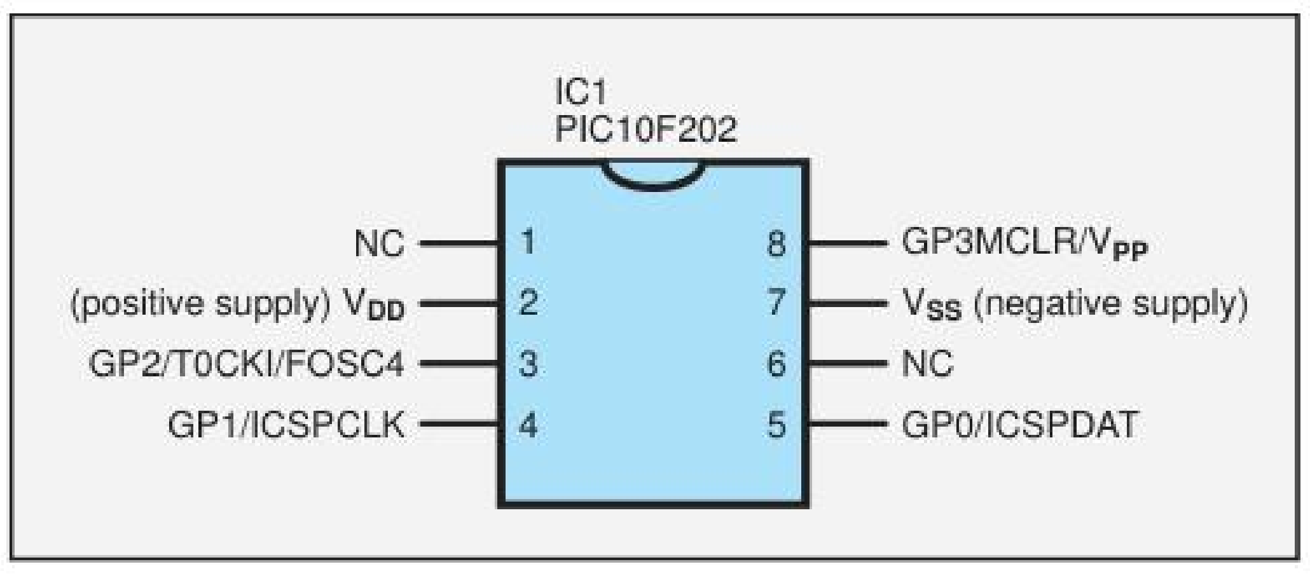


Fig.3. PIC10F202 pinout

In this series of articles we will explore several of these ideas, and other if suggested by readers – just pop your thoughts into the magazine chat forum at: **www.eeweb.com/forum/category/epe-magazine**

We will explore circuits from breadboard through to surface-mount PCB designs to show the range of applications.

Although these devices have at a maximum only 512 words of program memory, each of these words is 12-bits long and can hold a complete assembly level instruction, including any data associated with it. All instructions operate in a single instruction cycle (except when the instruction causes a branch or a jump in the program, in which case it takes two cycles). An instruction cycle takes 1µs, so that's a million instructions executed every second – fast enough for the simple functions that we might put this device to.

The trick with using such a limited device is applying it to projects that can accommodate it – if you find yourself having to add many external components (or more expensive components) to circumvent the processor's limitations, then you will probably be better off choosing a more powerful processor. We will choose our projects carefully to be relevant to a processor of this size.

The next challenge is what can I do with just 512 program instructions? A surprising amount, as it turns out. Even more surprising is that you do not have to revert to programming the chip in assembly language – we can write in the 'C' high-level language in many cases. There are some subtleties required in setting up the project code build files, but we will explain these issues and how to avoid them next month.

Software set-up

If you wish to follow along at home and build these circuits, you will need to download the software development tools from Microchip. We will be using the MPLAB X IDE, and the MPLAB XC Compilers. Both of these are provided for free by Microchip. MPLAB X IDE can be downloaded from here: **http://bit.ly/pe-may19-mc1**

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- Smallest and lightest
- 7 in 1: Oscilloscope, FFT, X/Y, Recorder, Logic Analyzer, Protocol decoder, Signal generator

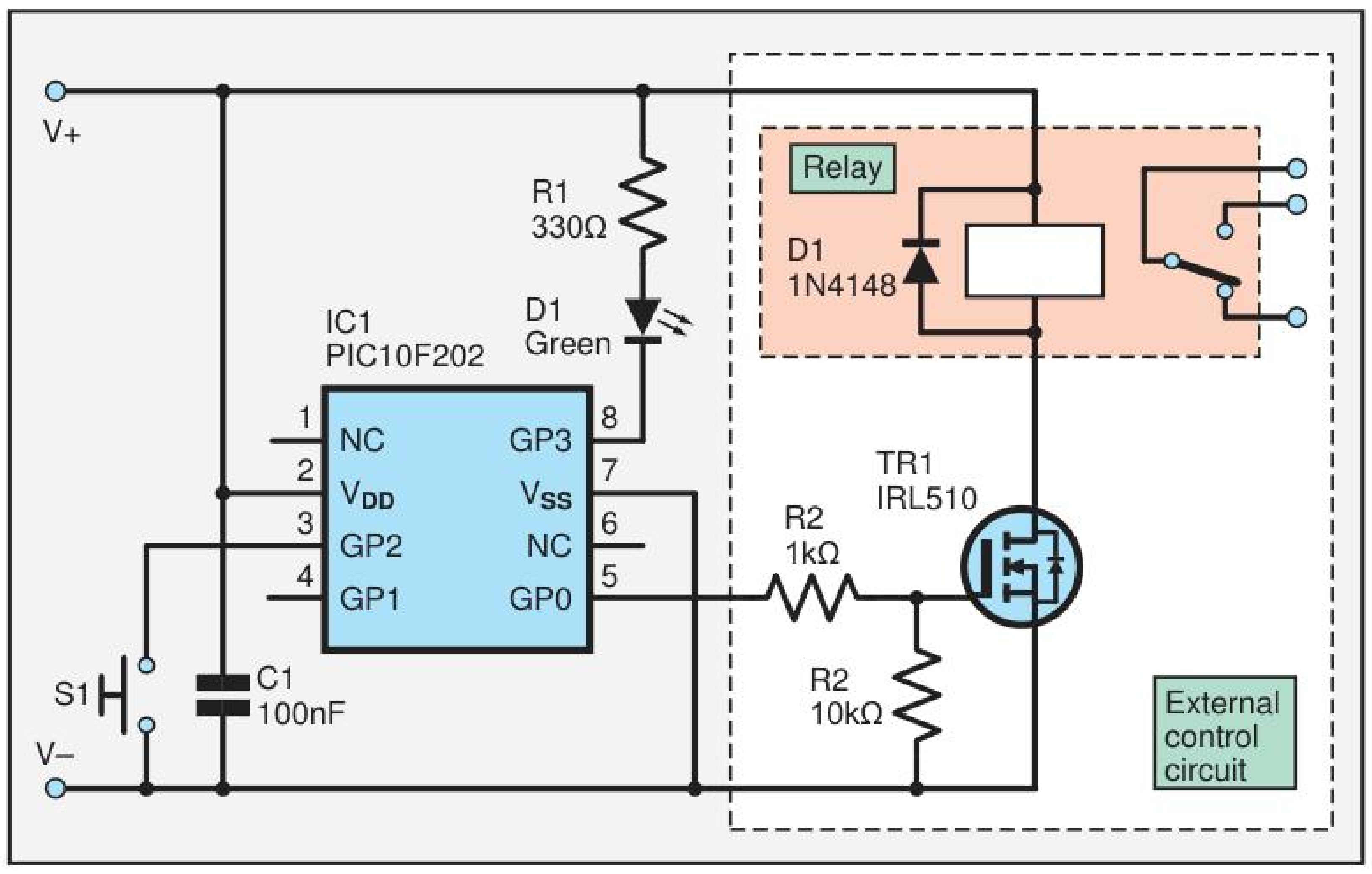


Fig.4. Our first PIC10F202 project schematic

The current version of MPLAB X IDE is v5.15, other versions close to this should be compatible with the project source files we will provide on-line. Installation is straightforward, and typically you accept all the default settings offered. Once the IDE installation completes and you have clicked 'Finish', your web browser should open several webpages on the Microchip website. Select the page titled 'MPLAB- XC Compliers' and select the 'MPLAB XC8 Complier' download. Installation is again straightforward, accepting the default option presented. Once completed, installation of the compiler will be automatically incorporated into MPLAB IDE – there are no additional set-up steps required.

Besides the software installation you will also require a PICkit 3 or PICkit 4 debugger device to load code onto an actual processor. Bear in mind that on a device of this size, debugging software (i.e., single-stepping through each line of execution) is not possible without an optional adaptor board. As our programs are going to be very small, and very simple, we have decided not to worry with this additional cost.

Do download the datasheet for the PIC10F200 family, which can be found here: **http://bit.ly/pe-may19-mc2**

Power of C

The assembly language instruction set of the PIC10F is very similar to other PIC processors you may be familiar with, including the PIC16 and PIC18 products. For the most part, however, this will not be relevant, as we will be developing our software in the higher-level programming language 'C'. We will demonstrate how efficient the 'C' compiler is, and what tricks we can apply to improve that efficiency when the compiler converts our code into assembly language. While there are only 512 words of Flash memory, the efficiency of the PIC microcontroller architecture means each one of those words can hold a complete single assembly instruction. Most instructions occupy a single word of memory. 'C' language program statements can occupy between one and six instructions typically, depending on the complexity of the particular instruction.

Our first project

For our first project we will attempt to create the 'smart power switch' described above. The circuit for this is shown in Fig.4. A relay will be incorporated to allow switching of higher voltage circuits, but we recommend that you do not attempt to do so on Veroboard or breadboard. The IRL510 FET is powerful enough to switch 1A loads directly, so the relay may well be optional. What this circuit could be used for we'll leave to your imagination! For lower current loads the FET could be replaced by a simple NPN transistor.

In summary

If you think this device is only for novelty functions, think again. It can provide valuable services as a sub-set of a larger circuit (just as the venerable NE555 timer chip has done for decades.) The author has just completed a commercial project where the PIC10F200 was the central microcontroller and has been deployed in the thousands. In that example, the code space used by the application was just 80 words, 35% of the available space, all written in 'C'. This is a device that is definitely worth taking seriously.

While these parts are readily available from the usual electronics distributors, such as Farnell and RS Components, delivery charges may be significantly higher than the actual device costs – these chips cost as little as 50 pence. eBay and Amazon can be a good source, however, just take care to check on the postage and packaging fees. Also, be careful to order the correct part – the plastic DIL package version is PIC10F202-I/P, while the six-pin surface mount part is PIC10F202T-I/OT.

Next month

In *Part 2*, next month we will look at how the 'C' compiler within MPLAB can be configured to reduce the code space used, build the first circuit and explore the code required to bring it to life.

Max's Cool Beans

By Max the Magnificent

The glorious Maxfield-Farr 4-bit HRRG Computer

I have to say that I'm super excited at the moment, because a rather cool hobby project that has been dormant for over ten years recently sprung back to life. This is something I'm working on with my chum, Joe Farr, who is a whizz at both hardware and software design.

The project in question involves designing and building a 4-bit computer from the ground up. Seriously, we're going to document how to implement this design using only primitive logic functions (eg, AND, OR, NAND, NOT), if anyone were to desire to go this far.

But this is not just any old 4-bit computer. No, indeedy-doody. This is the *Maxfield-Farr 4-bit HRRG Computer*, where HRRG stands for Heath Robinson and Rube Goldberg. As you are nodoubt aware, William Heath Robinson (1872-1944) and Reuben Lucius Goldberg (1883-1970) were both famous for creating illustrations of machines that were intended to perform relatively simple tasks, but whose implementations were incredibly complex such that they performed their chores in exceedingly convoluted and indirect ways.

So, why do we qualify our computer as being 'HRRG'? Well, let me elucidate (don't worry, I'm a professional).

Mixing technologies

This all started deep in the mists of time when I ran across a mega-cool electromechanical relay-based computer that was created by Harry Porter (no relation to Harry Potter; **http://bit.ly/pe-may19-cb1**). Harry's machine is gloriously presented in a series of glass-fronted wooden cabinets mounted to the wall.

I've long wanted to build a relay-based computer of my own, but I don't want to simply replicate something that someone else has already done. Thus was born the idea of a mixed-technology computer, in which the contents of each of the cabinets is implemented using a different technology, including (but not limited to) relays, vacuum tubes, discrete transistors, 74-series TTL logic, along with mechanical, magnetic, pneumatic, and hydraulic logic.

One thing with which I'm not enamoured about Harry's implementation is that his cabinets are linked by great big bundles of wires. I can't see my wife (Gina the Gorgeous) going for that, so the cabinets in our realisation will communicate via a wireless mesh network. Now, you may be saying to yourself, 'If this doesn't count as HRRG, I don't know what does,' to which I would reply, 'You ain't seen nothing yet!'

Decisions… decisions…

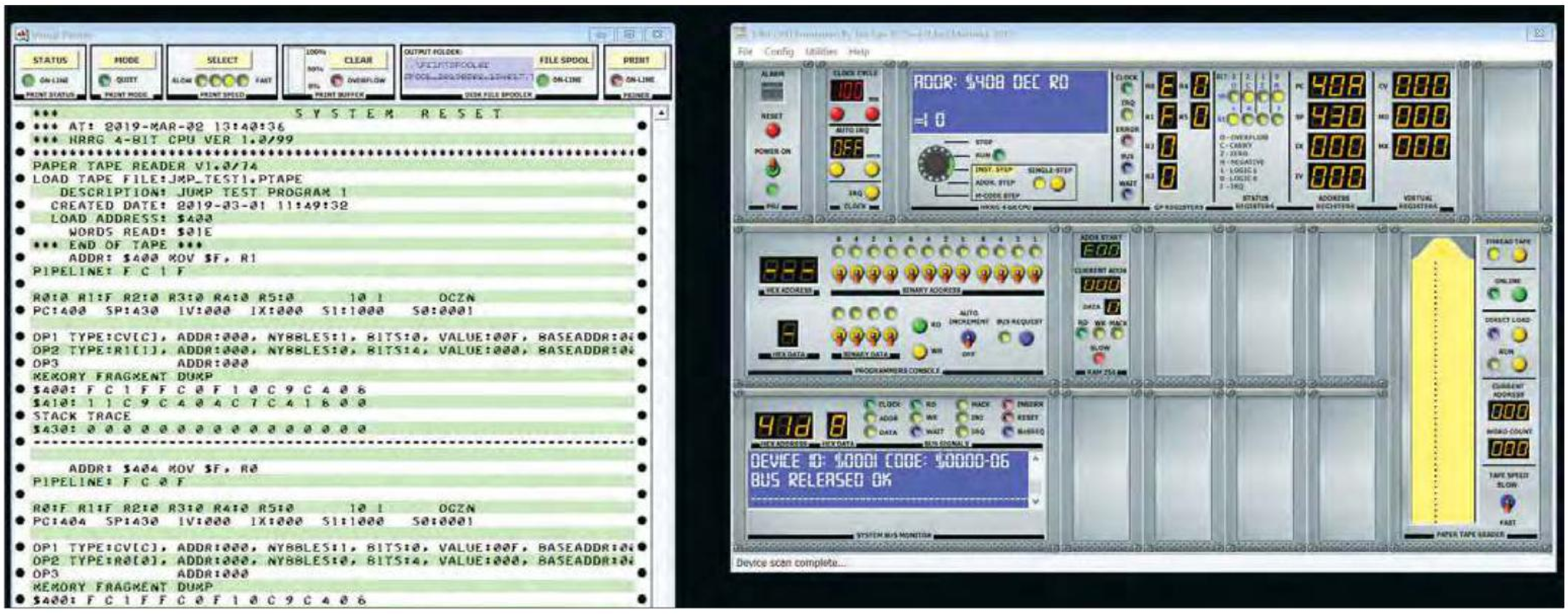
I tell you, you have no idea how many decisions are involved in this sort of project until you start writing things down. Fortunately, things are a lot easier where there are two of you to bounce ideas off each other. On the one hand, Joe and I were tempted to build an 8-bit computer, but we eventually opted for the 4-bit version for a number of reasons – not least because we want it to play an educational role, and – in some respects – a 4-bit machine will be easier for beginners to wrap their brains around.

Another consideration is that we find it fun to work around the limitations imposed by a 4-bit architecture, such as the fact that we have only 2^4 = 16 instructions to play with. Similarly, the fact that we have opted for a 12-bit address bus that supports only 2^{12} = 4,096 memory locations adds to the fun and frivolity.

Some of the early microprocessors had a single accumulator as their main internal data register. Others sported two accumulators or a bunch of general-purpose registers. Also, there is a range of possibilities when it comes to the addressing modes one decides to support, because each mode adds to the complexity of the instruction decode and addressing logic. Happily, after countless hours of discussion, we eventually tied everything down, now all we have to do is build the beast.

Returning to Earth

I'm not as stupid as I look, but there again, who could be? Remember that we want our computer to be educational, but even I realise that building


STATUS
ON-LINE
PRINT STATUS
MODE
QUIET
PRINT MODE
SELECT
SLOW
FAST
PRINT SPEED
100%
50%
0%
CLEAR
OVERFLOW
PRINT BUFFER
OUTPUT FOLDER:
FILE SPOOL
ON-LINE
DISK FILE SPOOLER
PRINT
ON-LINE
PRINTER
File
Config
Utilities
Help
ALARM
RESET
POWER ON
PSU
CLOCK CYCLE
100
AUTO IRQ
OFF
IRQ
CLOCK
STOP
RUN
INST. STEP
ADDR. STEP
M-CODE STEP
SINGLE-STEP
HRRG 4-bit CPU
CLOCK
IRQ
ERROR
BUS
WAIT
GP REGISTERS
STATUS REGISTERS
ADDRESS REGISTERS
VIRTUAL REGISTERS
HEX ADDRESS
BINARY ADDRESS
HEX DATA
BINARY DATA
RD
WR
AUTO INCREMENT
BUS REQUEST
OFF
PROGRAMMERS CONSOLE
ADDR START
CURRENT ADDR
DATA
SLOW
RAM 256
CLOCK
ADDR
DATA
RD
WR
WAIT
MACK
IRQ
INERRR
RESET
BUSREQ
HEX ADDRESS
HEX DATA
BUS SIGNALS
SYSTEM BUS MONITOR
PAPER TAPE READER
ON-LINE
DIRECT LOAD
RUN
CURRENT ADDRESS
WORD COUNT
TAPE SPEED
SLOW
FAST

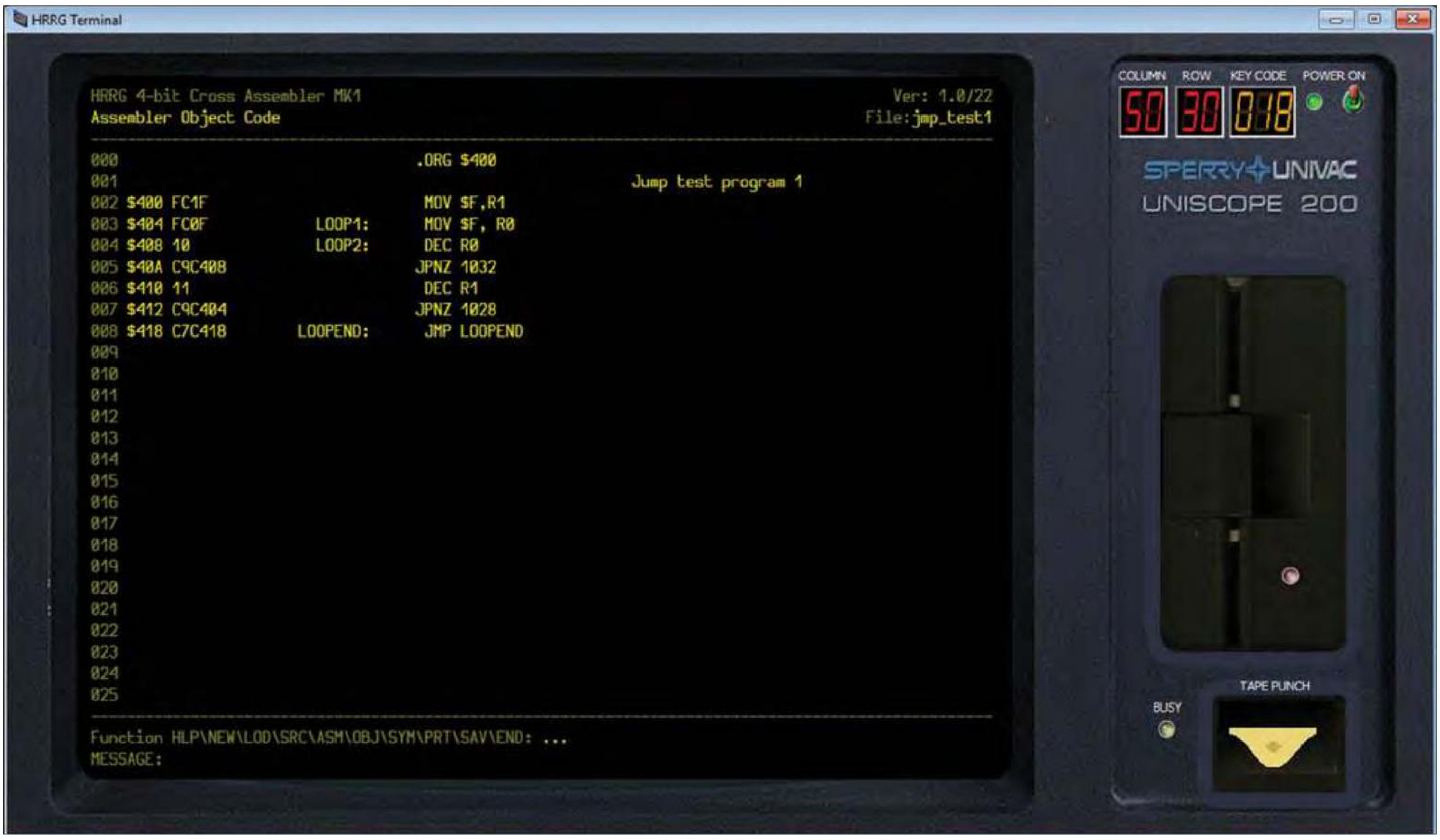
```
***                               S Y S T E M   R E S E T
*** AT: 2019-MAR-02 13:40:36
*** HRRG 4-BIT CPU VER 1.0/99
*********************************************************************
PAPER TAPE READER V1.0/74
LOAD TAPE FILE:JMP_TEST1.PTAPE
   DESCRIPTION: JUMP TEST PROGRAM 1
  CREATED DATE: 2019-03-01 11:49:32
  LOAD ADDRESS: $400
    WORDS READ: $01E
*** END OF TAPE ***
    ADDR: $400 MOV $F, R1
PIPELINE: F C 1 F

R0:0 R1:F R2:0 R3:0 R4:0 R5:0        10 1        OCZN
PC:400  SP:430  IV:000  IX:000  S1:1000   S0:0001

OP1 TYPE:CV[C], ADDR:000, NYBBLES:1, BITS:0, VALUE:00F, BASEADDR:0
OP2 TYPE:R1[1], ADDR:000, NYBBLES:0, BITS:4, VALUE:000, BASEADDR:0
OP3             ADDR:000
MEMORY FRAGMENT DUMP
$400: F C 1 F F C 0 F 1 0 C 9 C 4 0 6
$410: 1 1 C 9 C 4 0 4 C 7 C 4 1 6 0 0
STACK TRACE
$430: 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 0
---------------------------------------------------------------------

    ADDR: $404 MOV $F, R0
PIPELINE: F C 0 F

R0:F R1:F R2:0 R3:0 R4:0 R5:0        10 1        OCZN
PC:404  SP:430  IV:000  IX:000  S1:1000   S0:0001

OP1 TYPE:CV[C], ADDR:000, NYBBLES:1, BITS:0, VALUE:00F, BASEADDR:0
OP2 TYPE:R0[0], ADDR:000, NYBBLES:0, BITS:4, VALUE:000, BASEADDR:0
OP3             ADDR:000
MEMORY FRAGMENT DUMP
$400: F C 1 F F C 0 F 1 0 C 9 C 4 0 6
```

```
ADDR: $408 DEC R0
=1 0
```

```
DEVICE ID: $0001 CODE: $0000-06
BUS RELEASED OK
```

Device scan complete...

Fig.1. The virtual printer (left) and 4-Bit HRRG Emulator (right).



```
HRRG 4-bit Cross Assembler MK1                                        Ver: 1.0/22
Assembler Object Code                                              File:jmp_test1
---------------------------------------------------------------------------------
                                .ORG $400
                                                   Jump test program 1
$400 FC1F                        MOV $F,R1
$404 FC0F          LOOP1:        MOV $F, R0
$408 10            LOOP2:        DEC R0
$40A C9C408                     JPNZ 1032
$410 11                          DEC R1
$412 C9C404                     JPNZ 1028
$418 C7C418      LOOPEND:        JMP LOOPEND
---------------------------------------------------------------------------------
Function HLP\NEW\LOD\SRC\ASM\OBJ\SYM\PRT\SAV\END: ...
MESSAGE:
```

Fig.2. The HRRG Terminal (Readers of a certain vintage will recall the Sperry Uniscope was a class of computer terminals from the 1960s).

a full-up mixed-technology 4-bit HRRG will be a costly and time-consuming task.

Thus, in addition to the physical realisation, we are also creating a virtual version of the machine – called the *HRRG Emulator*. This is the part that Joe is currently working on (Fig.1). The virtual printer shows the system being reset, a test program being loaded from tape, and then the audit trail of the test program being single-stepped (you can vary the level of information being printed or switch it off completely as required).

The emulator is presented in the form of a three-shelf rack. Each shelf contains a number of modules. In the upper shelf, from left to right, we see the main power control module, the system clock module, and the main CPU module. On the left-hand side of the middle shelf we see the Programmers Console, which allows you to enter programs by hand (well, mouse-click) using toggle switches and push buttons. On the lower right-hand side, we see the virtual paper-tape reader, which allows you to read in programs from virtual paper tapes.

Speaking of which, Fig.2 shows the virtual HRRG Terminal running our assembler. On the screen we see the object code that the assembler generated for this test program. This code was then punched to virtual paper tape, which was loaded into the emulator.

We will be adding additional modules, including ROM and RAM and a variety of input/output (I/O) devices, plus users will be able to add their own devices and configure which modules go where in the rack. Once we've ironed out all of the bugs, we will be making this virtual machine available for everyone to download and play with.

But wait, there's more… This is where things get really clever. Joe has architected the virtual machine to reflect the longer-term physical implementation, including the mechanisms by which the various modules (cabinets in the real world) talk to each other.

The idea is that you can start playing with the emulator running on a PC, then – when you are ready – you can commence creating one or more physical cabinets that can

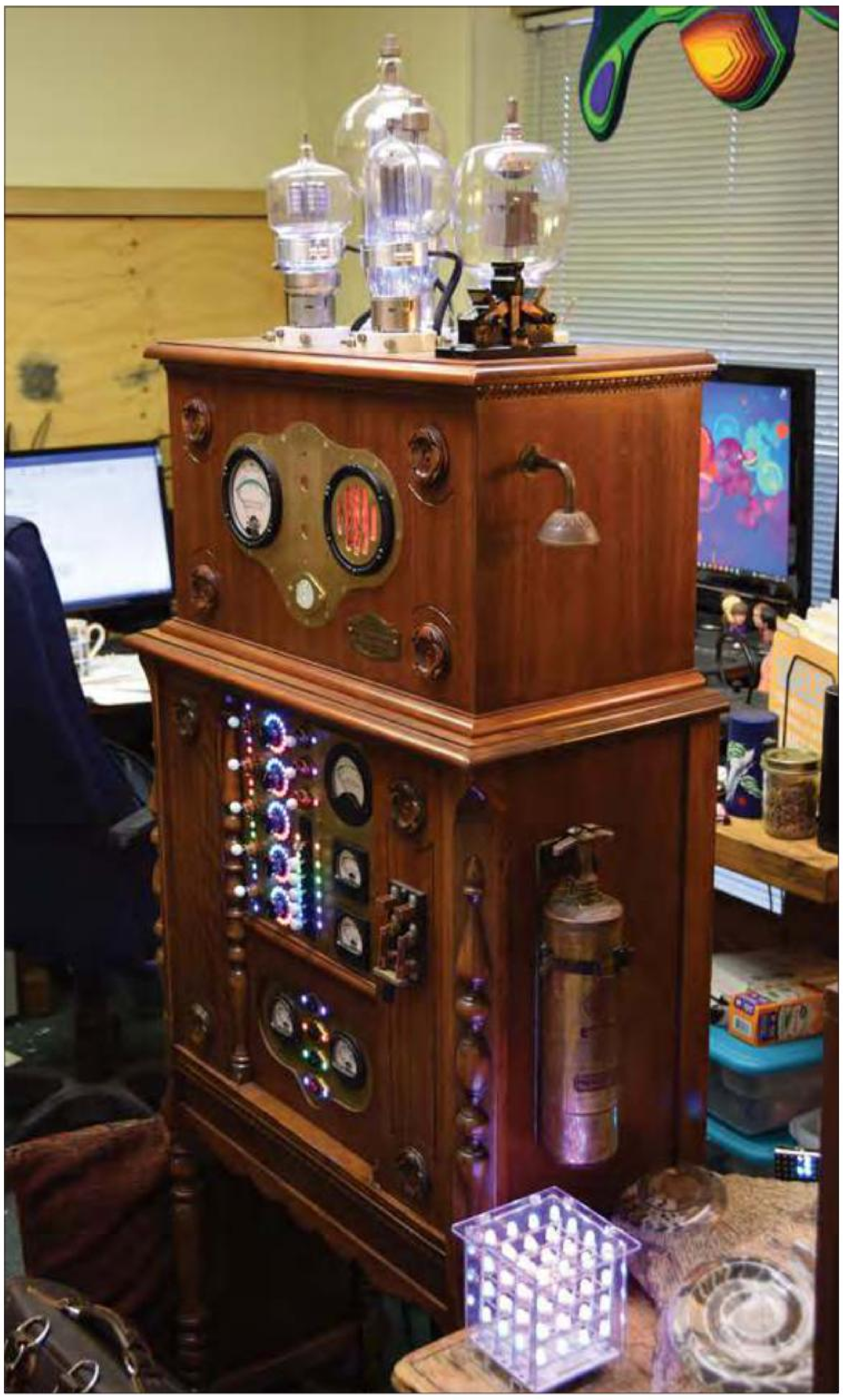


Fig.3. The current state of the *Inamorata Prognostication Engine.*

communicate (via wireless USB) with the emulator. Each of these physical cabinets can represent as much as the entire CPU, or as little as a single 4-bit word in memory.

This is where I think the real educational value of the *4-Bit HRRG Computer* begins to manifest itself. A high school class could build a cabinet containing a small 4 × 4 array of memory implemented in a technology of their choice, and then see their program use the contents of this cabinet to perform its calculations.

I think I'll give you a little time to ponder the implications of what we've discussed thus far before I tell you our ideas for connecting multiple virtual and physical cabinets via the Internet. If you are interested in learning more about all of this, follow my articles on **EEWeb.com** (**http://bit.ly/pe-may19-cb2**) and keep on reading my *Cool Beans* columns here in *Practical Electronics*.

The colour of prognostication

I must admit that I'm super-excited by all the interest and emails I've received regarding the *Awesome Audio-Reactive Artifact* project I introduced in the March 2019 issue of *Practical Electronics*. In fact, I recently displayed this little beauty at a Hamfest and electronics flea market in Birmingham, Alabama, US. You can see a video and access the code for this project in an article I posted on **EEWeb.com** (**http://bit.ly/pe-may19-cb3**).

I'm also amazed that people are still emailing me to inquire about the status of my *Inamorata Prognostication Engine* project that I discussed deep in the mists of time circa 2014 (where does the time go?). As you may recall, the purpose of the *Prognostication Engine* is to predict my wife's mood of the moment when I prepare to head home from my office at the end of the day. (Ironically, if she ever discovers what the *Prognostication Engine* is actually for, I'm reasonably confident that I won't need it to divine her disposition.)

In fact, this project – which currently has pride of place in my office – is starting to look very tasty; however, it's still a long way from completion. The reason this is taking so long is that I usually have multiple projects on the go at any particular time and I'm very easily distrac… *Squirrel!*

As we see in Fig.3, the vacuum tubes sitting on top of the engine are illuminated with tri-colour LEDs, as are the dials and switches in the lower part of the unit. Observe the inverted brass cone sticking out of the upper right-hand side of the cabinet. Also observe the antique brass and copper fire extinguisher attached to the side of the lower cabinet. The way we power the beast up is to hold a flame under the inverted brass cone. After a few seconds, we hear a 'pop' sound as this causes the furnace to ignite. This is the point where we have to furiously work a 100-year-old leather foot pump (that originally graced a dentist's office) to get the furnace roaring sufficiently hot that we can throw the master knife switch to let the automatic systems take over. You can see a video showing the *Prognostication Engine* in action in an article I posted on **EEWeb.com** (**http://bit.ly/pe-may19-cb4**). I will of course keep you appraised of any developments on all of my projects in future *Cool Beans* columns here in *Practical Electronics*.

Cool bean Max Maxfield (Hawaiian shirt, on the right) is editor-in-chief at **EEWeb.com** – the go-to site for users of electronic design tools and askers of electronic questions.

Comments or questions? Email Max at: **max@CliveMaxfield.com**

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Make it with Micromite

Phil Boyce – hands on with the mighty PIC-powered, BASIC microcontroller

Part 4: Digital inputs and outputs

For those of you following this series, by now you should have your *Micromite Keyring Computer* (*MKC*) communicating with a PC (using TeraTerm) via the plug-in *Development Module* (*DM*). If so, then you are nearly ready to begin exploring MMBASIC and start working through the topics discussed in the upcoming articles. However, if you've not yet had success for any reason, and you need help, then simply drop us an email and we will get you started.

This month, we will begin by performing the same kind of LED test discussed in *Part 2* (when checking the *MKC*) to finish testing the *DM*. This will ensure that all the input/output (I/O) pins on the *DM* are correctly connected. This is an essential step and should not be skipped.

We will then configure a couple of *MKC* options and change some settings in TeraTerm to make the development environment more pleasant to use.

Next, we will touch on the basics of MMBASIC, just to ensure that we are all aware of some fundamental concepts regarding the Micromite and MMBASIC.

After that, the fun will begin; starting with how to control a simple digital output (LED), and then how to monitor a digital input (button). If you're thinking this isn't very exciting, please don't worry – we will quickly expand our activities to demonstrate just how easy it is to use MMBASIC to add 'intelligence' to these basic components.

Ultimately, we will work towards our first mini-project, which will use the concepts that we'll be covering here. That mini-project, which will appear in next month's article, is an *Electronic Dice* – see Fig.1. Although it is just seven LEDs (digital outputs) and one button (digital input), please don't underestimate it. My – 'surely there's an app!' – children have explored many board games thanks to the fun they have had when using this *Electronic Dice*! And it will teach you some very useful Micromite basics.

Questions? Please email Phil at:
contactus@micromite.org

We mentioned in *Part 1* that the idea of this series is to give you exposure to MMBASIC and to make you aware of what is possible, so don't be afraid to modify the hardware and code that we present here, after all, that is the best way to learn!

It is assumed at this point that your *MKC* is successfully communicating with TeraTerm. If not, then please perform all the tests discussed in *Parts 2 & 3* before contacting us for help.

Testing the *DM*

Before you continue with any of the topics coming up, it is very important that you successfully perform the LED test on the input/output headers on the *DM*. Each of the 19 input/output pins has two points to test on the *DM*, hence 38 points in total will need checking. These points are all shown highlighted in Fig.2.

To perform the test, simply grab yourself an LED and a resistor with a value between 330Ω and 1kΩ. Connect them as shown in Fig.2, either by soldering them together, or by using a breadboard and some jumper wires. Ensure that the LED's shorter lead is properly and securely inserted into the 4-way 0V header J7 (any of the four points is suitable). The test probe can be poked into the socket point under test.

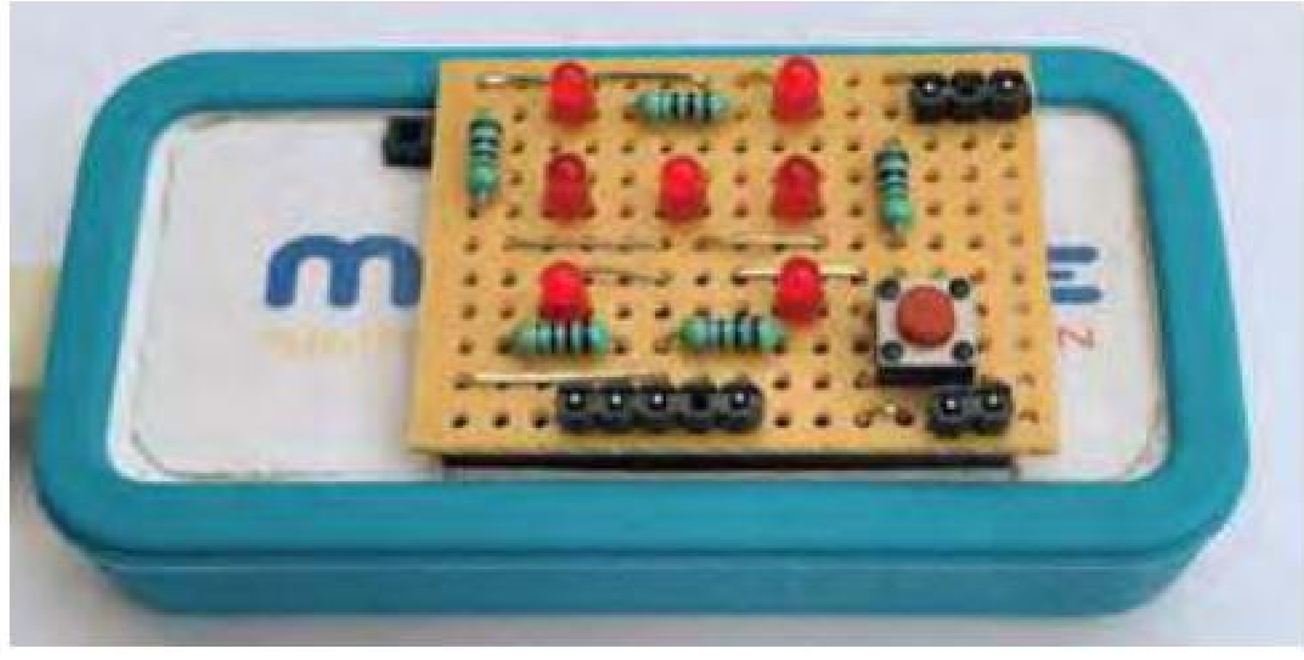


Fig.1. This *Electronic Dice* will be our first mini-project, appearing in next month's article.

The first thing to do is to test your actual test-circuit! So plug the *DM* into the *MKC* and power it up (ensure the two power LEDs come on). Connecting to TeraTerm will help in a little while, so go ahead and do that too (press the *DM* reset button and ensure you see the Micromite start up message – remember to set baud speed to 38400, as discussed in *Part 3*). Now insert the test probe into any position on the middle 4-way header, J8. The four points on J8 are all at 3V3, so your test LED should turn on. If it does, then great; however, if the test LED does not light up then you will need to check a few things.

Begin by ensuring the *MKC* and *DM* are powered up (check power LEDs). Check you have the LED in the correct orientation

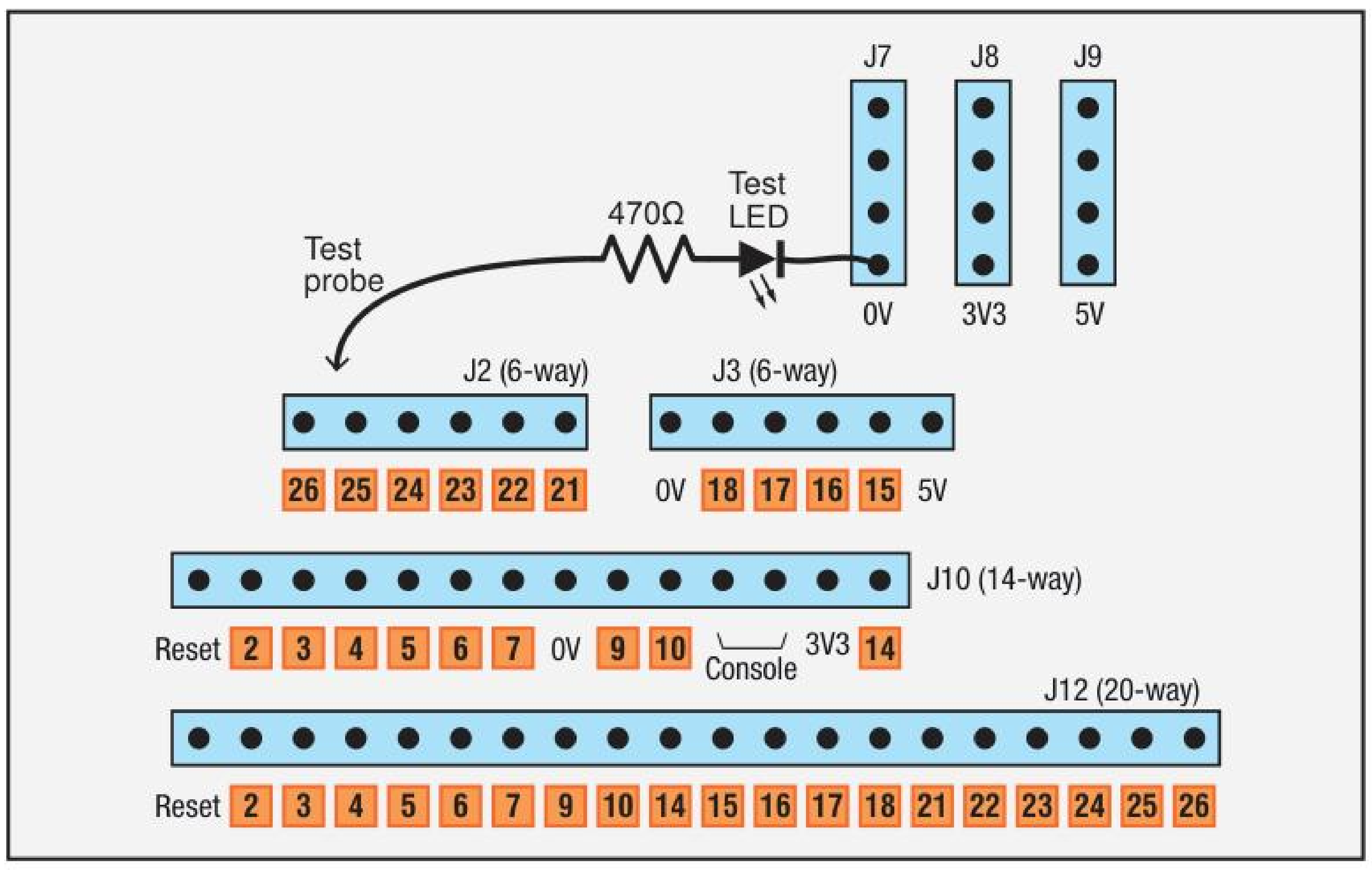
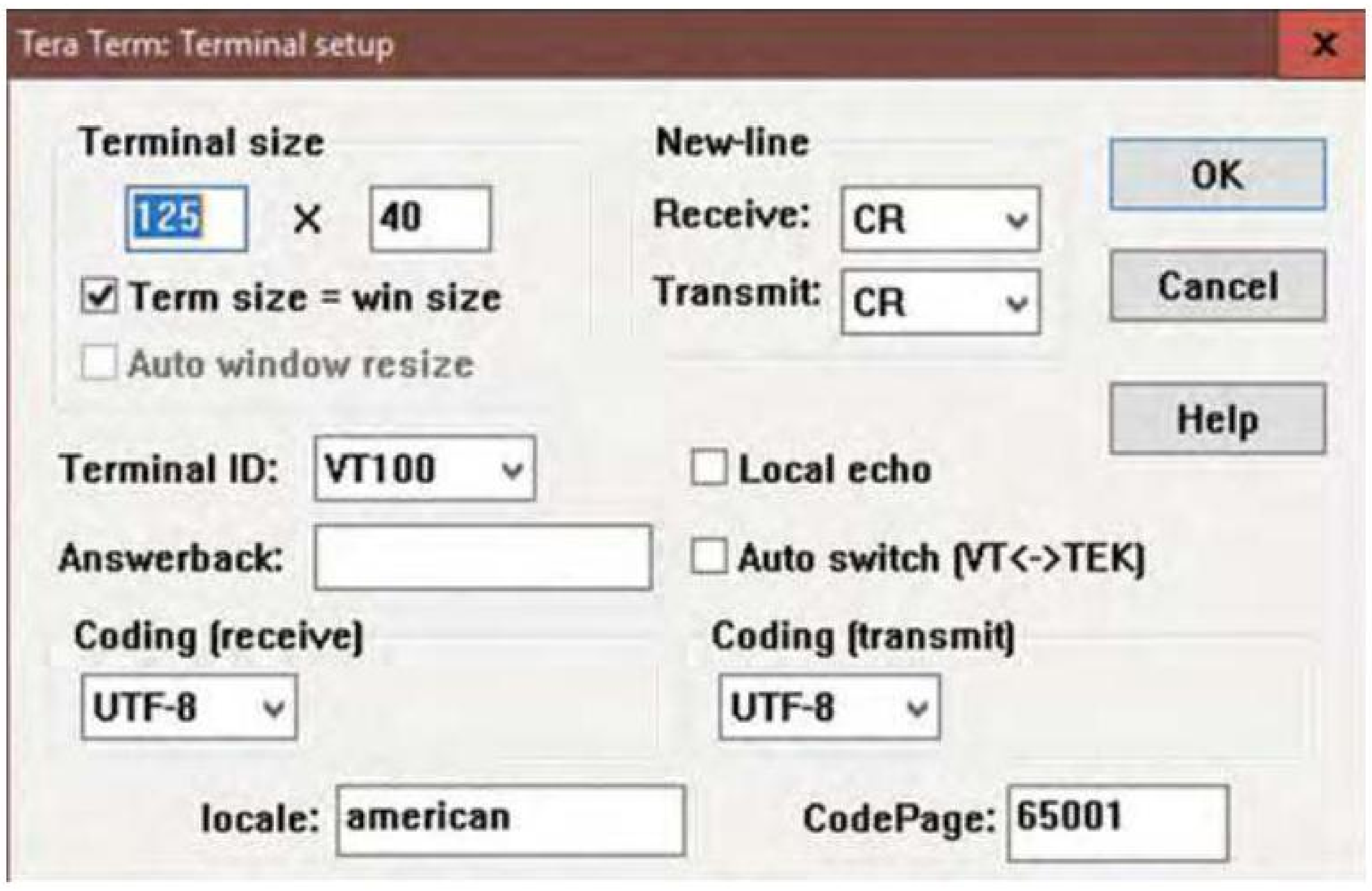


Fig.2. Summary of the 19 I/O pin numbers and locations on the *Development Module* (*DM*). Also shown is the simple test circuit.


Tera Term: Terminal setup
OK
Cancel
Help

Terminal size: 125 × 40

☑ Term size = win size

☐ Auto window resize

New-line: Receive: CR; Transmit: CR

Terminal ID: VT100

☐ Local echo

Answerback:

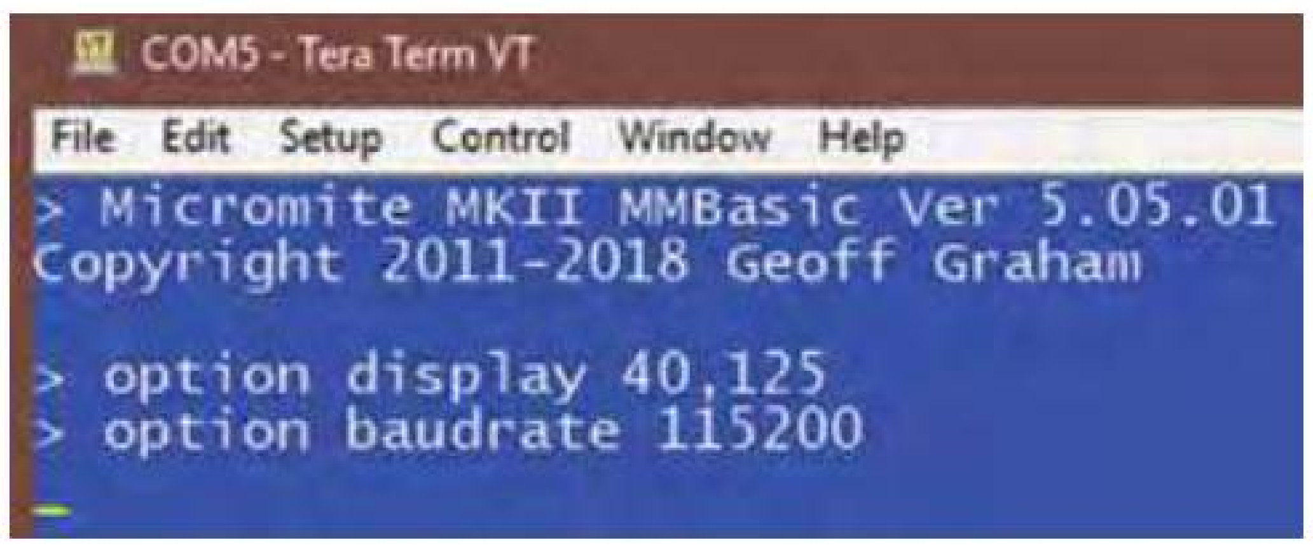
☐ Auto switch (VT<->TEK)

Coding (receive): UTF-8

Coding (transmit): UTF-8

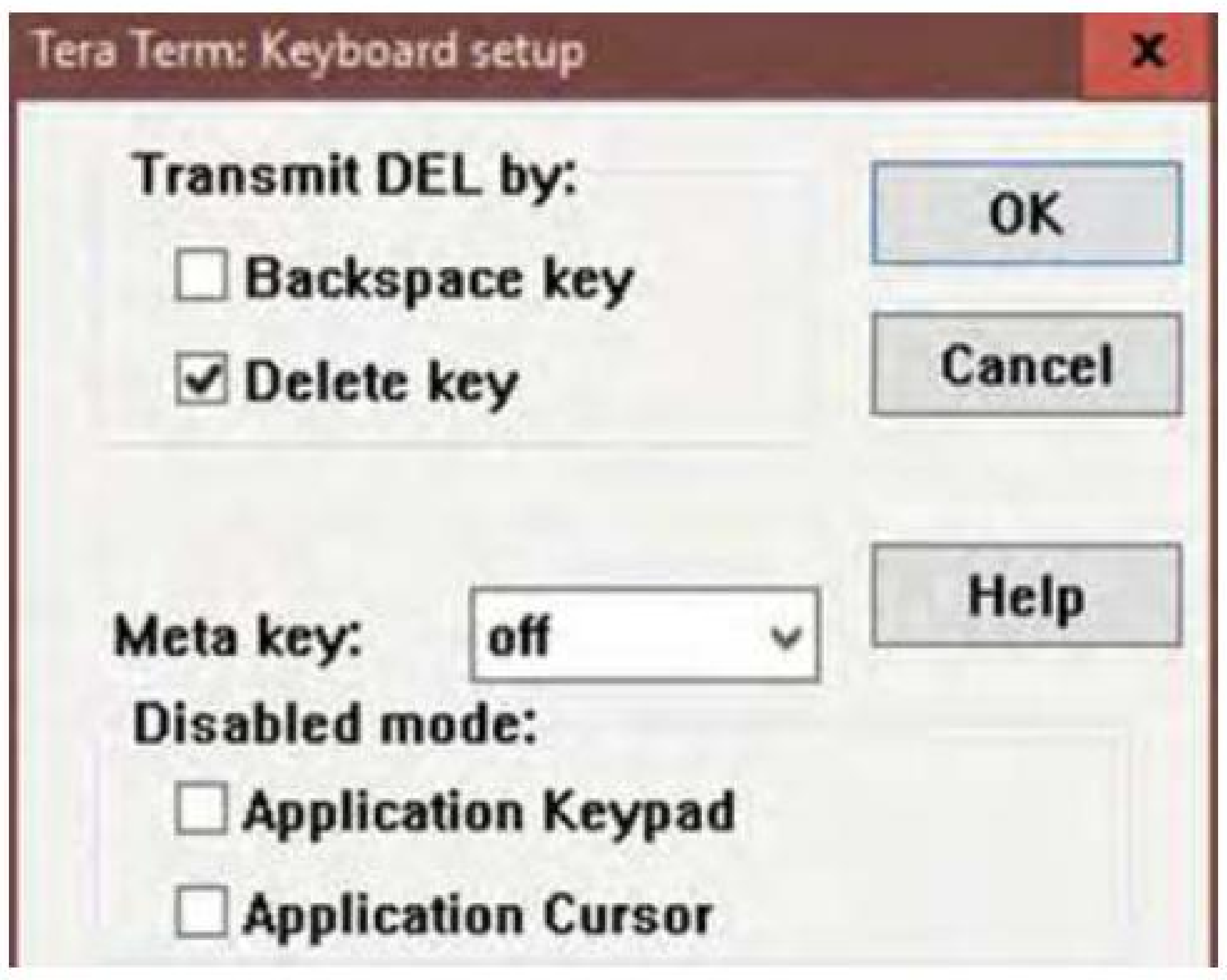
locale: american

CodePage: 65001


COM5 - Tera Term VT
File
Edit
Setup
Control
Window
Help

```
> Micromite MKII MMBasic Ver 5.05.01
Copyright 2011-2018 Geoff Graham

> option display 40,125
> option baudrate 115200
```


Tera Term: Keyboard setup
OK
Cancel
Help

Transmit DEL by:

☐ Backspace key

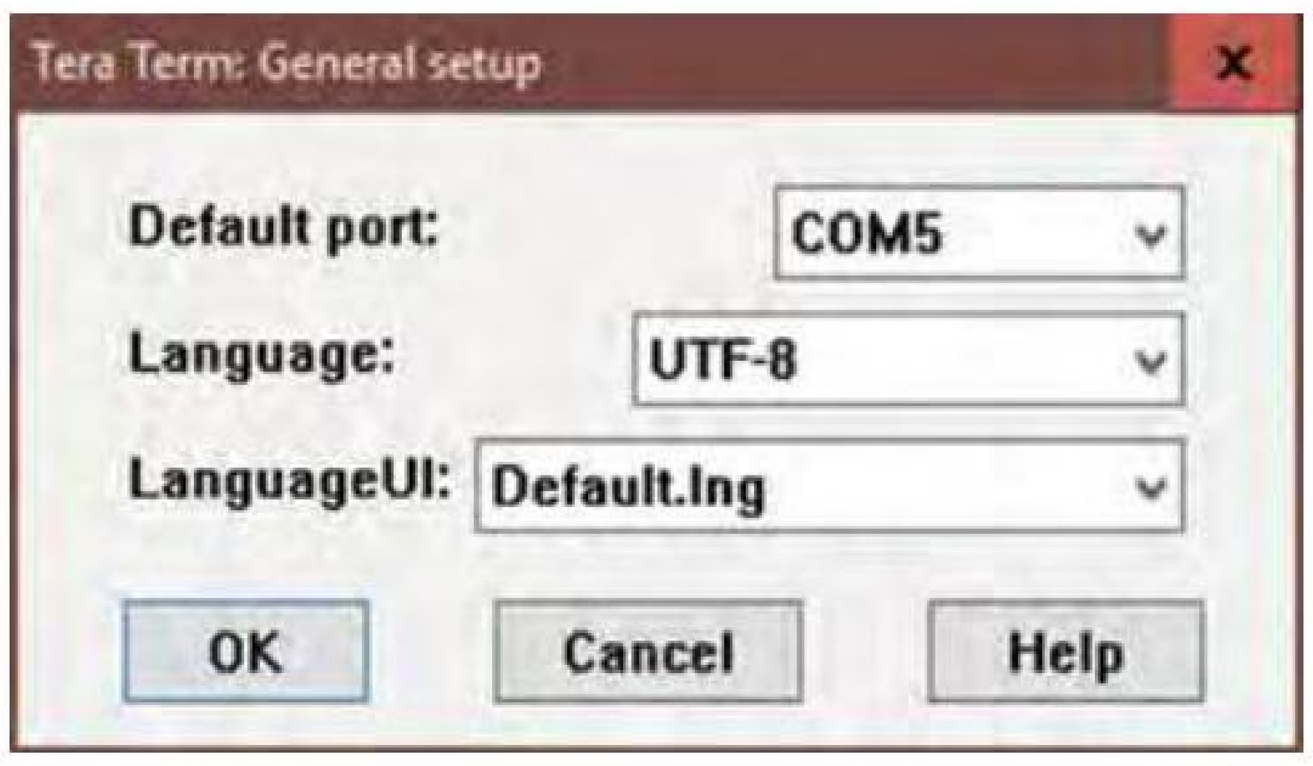
☑ Delete key

Meta key: off

Disabled mode:

☐ Application Keypad

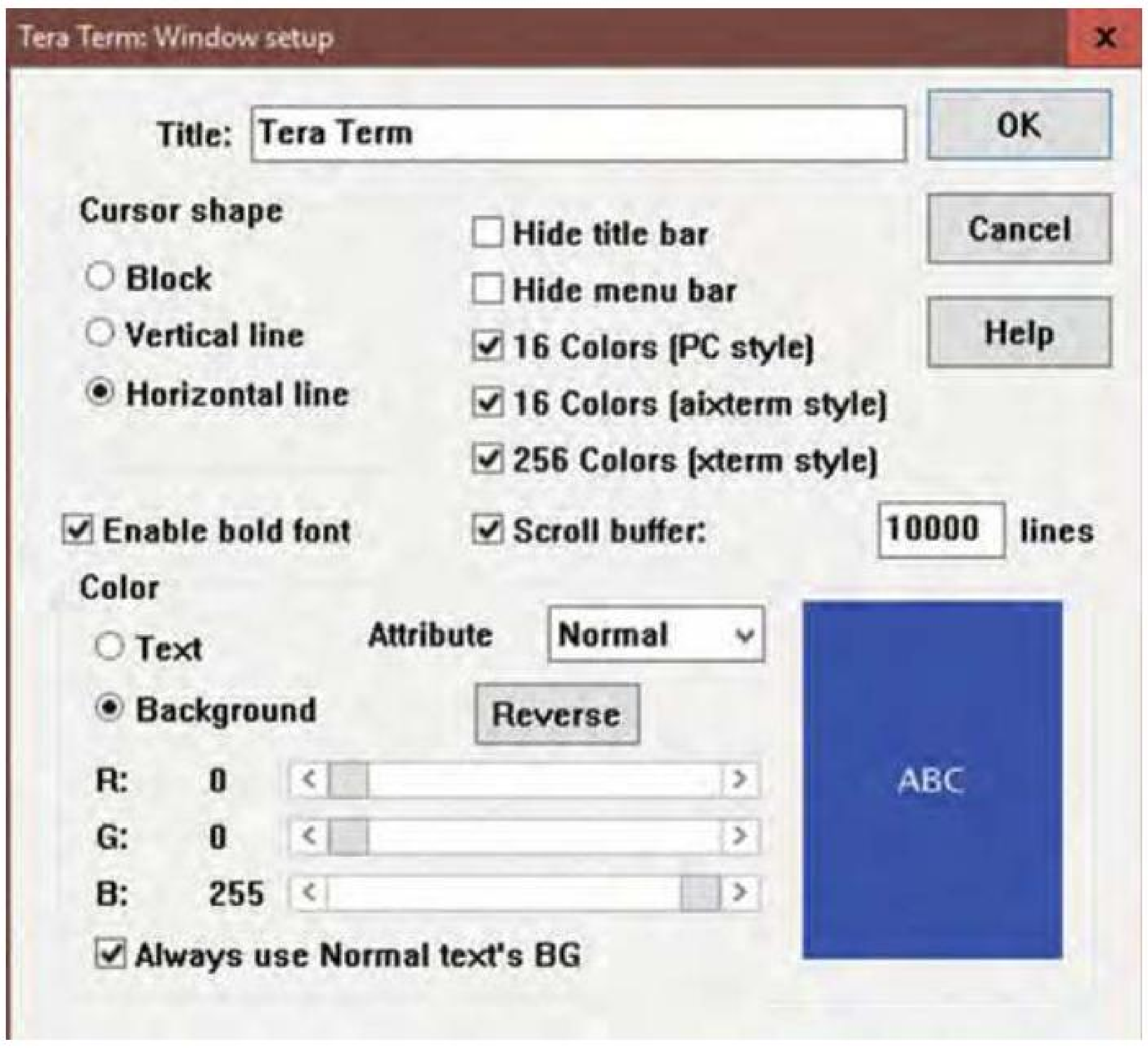
☐ Application Cursor


Tera Term: General setup
OK
Cancel
Help

Default port: COM5

Language: UTF-8

LanguageUI: Default.lng


Tera Term: Window setup
OK
Cancel
Help
Reverse
ABC

Title: Tera Term

Cursor shape: ○ Block ○ Vertical line ● Horizontal line

☐ Hide title bar

☐ Hide menu bar

☑ 16 Colors (PC style)

☑ 16 Colors (aixterm style)

☑ 256 Colors (xterm style)

☑ Enable bold font

☑ Scroll buffer: 10000 lines

Color: ○ Text ● Background

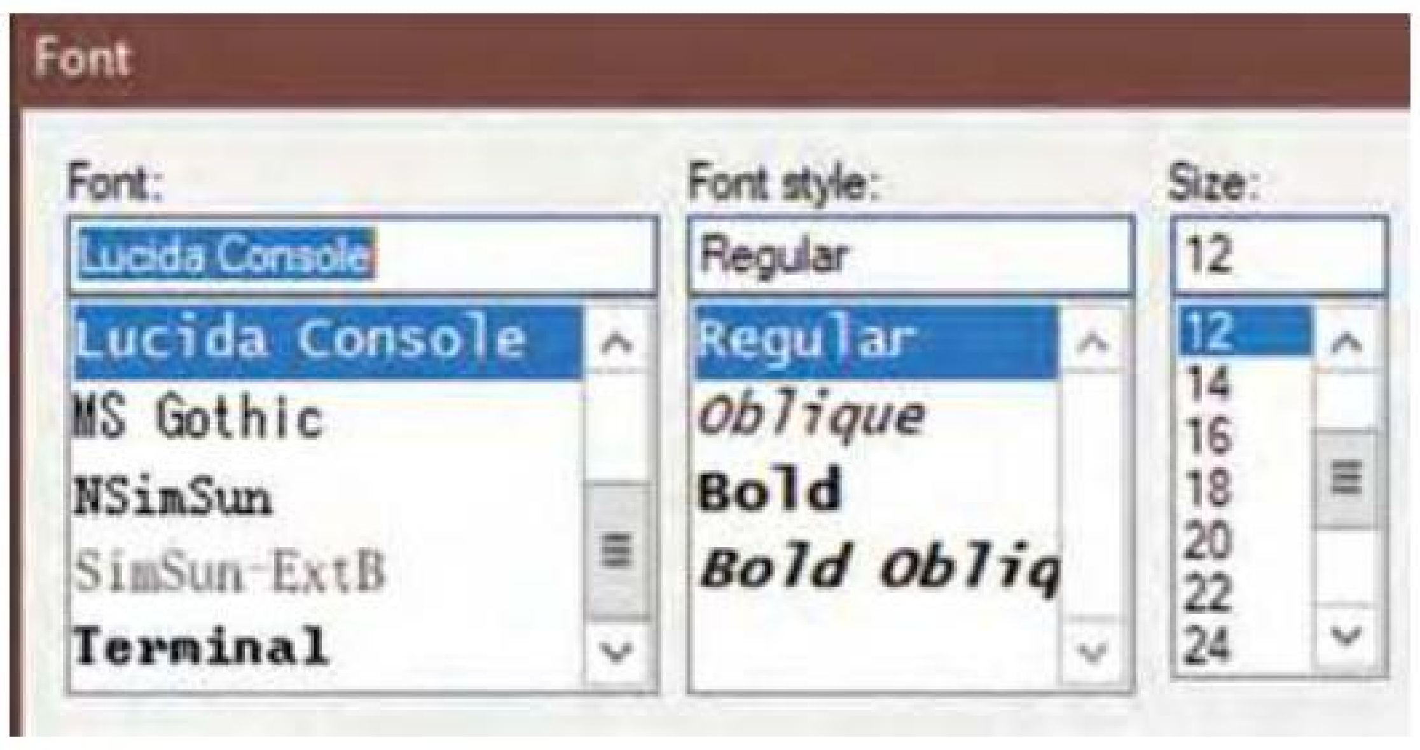
Attribute: Normal

R: 0

G: 0

B: 255

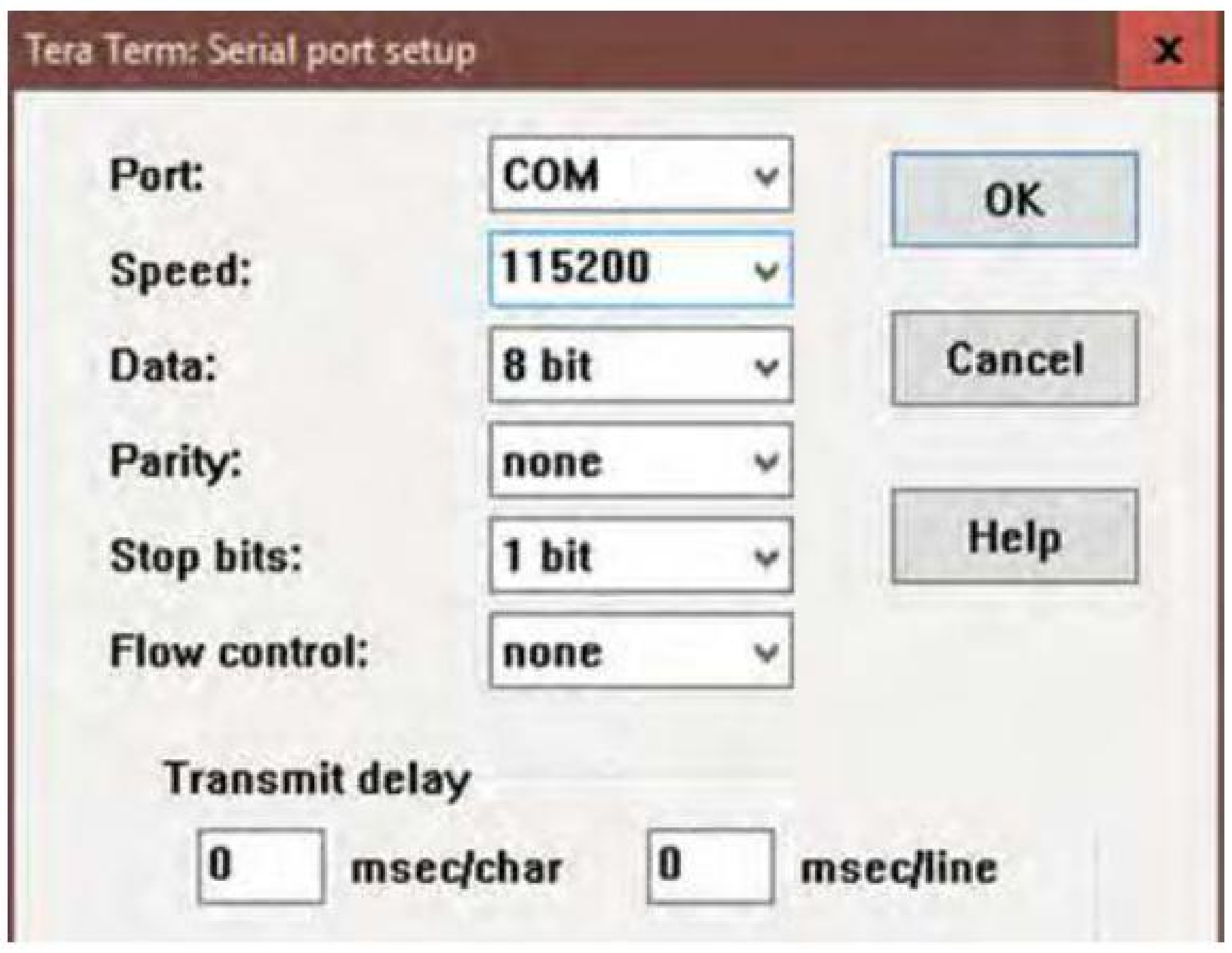
☑ Always use Normal text's BG


Font
Lucida Console
MS Gothic
NSimSun
SimSun-ExtB
Terminal
Regular
Oblique
Bold
Bold Obliq
12
14
16
18
20
22
24

Font: Lucida Console

Font style: Regular

Size: 12


Tera Term: Serial port setup
OK
Cancel
Help

Port: COM

Speed: 115200

Data: 8 bit

Parity: none

Stop bits: 1 bit

Flow control: none

Transmit delay: 0 msec/char 0 msec/line

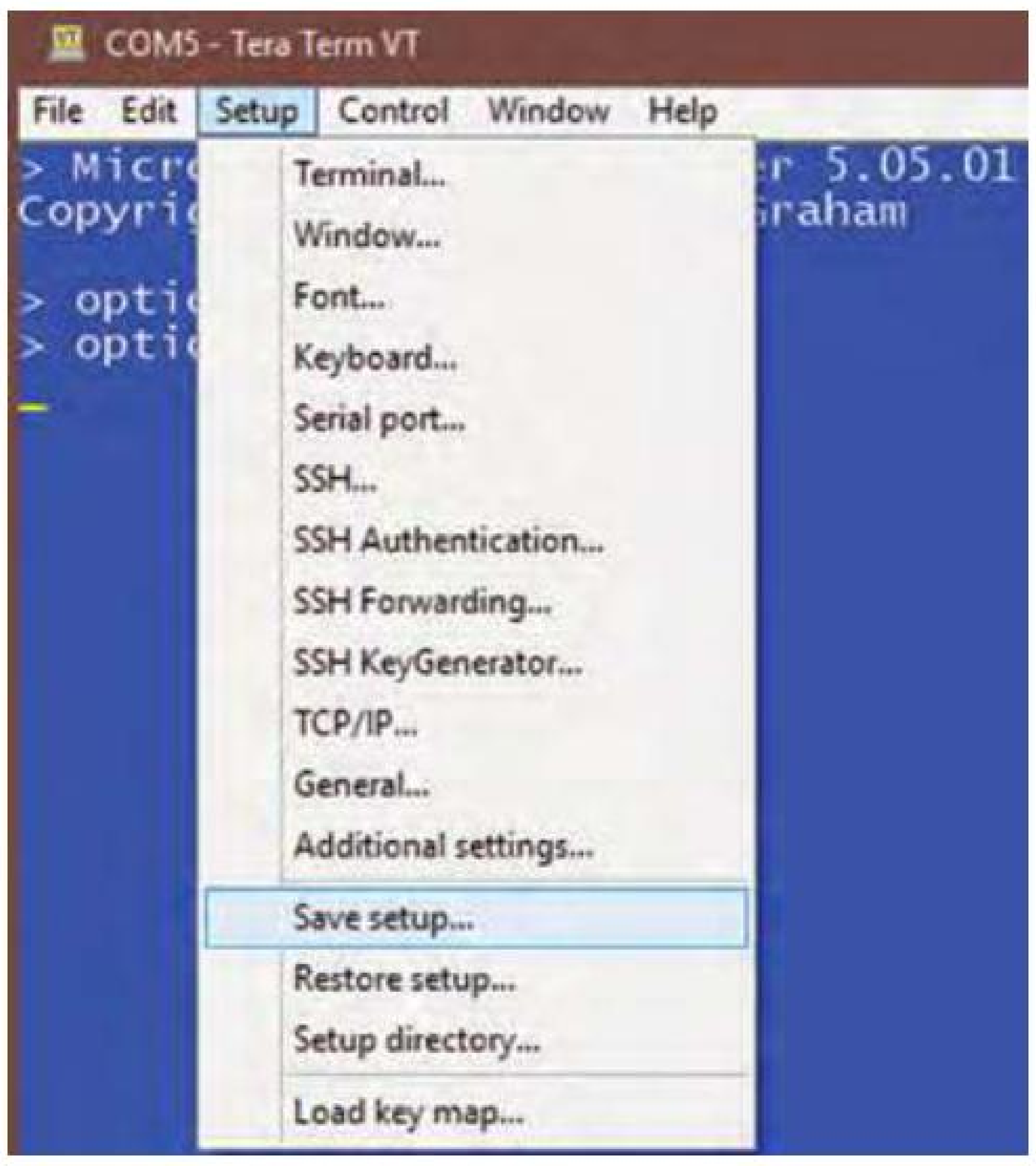

COM5 - Tera Term VT
File
Edit
Setup
Control
Window
Help
Terminal...
Window...
Font...
Keyboard...
Serial port...
SSH...
SSH Authentication...
SSH Forwarding...
SSH KeyGenerator...
TCP/IP...
General...
Additional settings...
Save setup...
Restore setup...
Setup directory...
Load key map...

Fig.3. Ensure your TeraTerm setup screens match these settings and remember to save!

– remember it is a polarised component; the short lead (k) needs to be connected to 0V. Also check the resistor value. If you have a value much higher than 1kΩ then the LED may be too dim to see if you're working in bright sunlight. These quick checks will resolve any issue and your test circuit is now ready to be used for testing the I/O points on the *DM*.

Errata

In Part 3 (Fig.2 on page 63) the component side of the stripboard layout incorrectly shows a track cut at position X21. This cut should not be there. The track side of the stripboard in Fig.2 is correctly shown with no cut at X21. If you have made this cut, then simply solder a small wire link to bridge the cut. It is mentioned here because a cut at X21 would isolate I/O pin 15 from the 20-way header; and hence the following test would fail for I/O 15 on J12.

There are now three possible scenarios for your *MKC* test software – it is:

1. Installed and running
2. Installed but is not running
3. Not installed.

You may know the scenario you are in, but for now just follow the guide below.

Referring to Fig.2, insert the test probe into the position highlighted with number 2 on the 14-way header J10. If you see the test LED briefly flash at approximately one-second intervals, then you have the *MKC* test software running on your *MKC* (scenario 1).

If the test LED is not flashing, then type RUN on the TeraTerm keyboard and check to see if the word RUN appears on the TeraTerm screen. If so, press the Enter key and check to see if the test LED flashes. If the test LED begins to flash, then scenario 2 applies to you; otherwise it is scenario 3.

For scenarios 1 and 2, simply go ahead and work through the 19 numbered points on J10, J3 and J2. At each point you should see the test LED flash. If any point does not flash, then there is an issue with that I/O signal between the PIC inside the *MKC* and the point under test. Check for open circuits and for track shorts using the methods discussed in *Part 2*, page 69 (*Useful test info* section). For those of you that have installed the 20-way header (J12), go ahead and test the 19 I/O points on that too (note that the left-hand point (reset) will not flash the test LED). If everything checks out then jump over the next section and go to: *Configuring your MKC and TeraTerm*.

If you are here then you must fall under scenario 3 (ie, *MKC* test software not installed). Hopefully, if you type anything into TeraTerm you will see the characters appear on the TeraTerm screen. If not, then Press (and hold down) the Ctrl key and then press the C key. This will stop any program that may be running. Now press any key and you should see the character(s) appear on the TeraTerm screen. If not, then you will need to check your TeraTerm communication, as discussed in *Part 3*.

Once you have the command prompt on the TeraTerm screen (the '>' character) and are able to type and see the characters, then you can begin to methodically test the 19 I/O signals on the *DM* as follows.

Insert the test probe into the first point (number 2) and type `SETPIN(2),DOUT` followed by the Enter key. Next type `PIN(2)=1` and again hit Enter. The test LED should now turn on (if not, first check for typing errors on the TeraTerm screen and also check your test probe is at point 2). If the LED is still off, then you have an issue with I/O number 2 somewhere between the *MKC* and the point being tested.

If your test LED successfully turned on, then move the test probe into the next point (number 3 in this case). The test LED should now be off. If however it is on, then you have a short somewhere between the tracks for I/O number 2 (which is on) and the tracks for I/O number 3 (which is 'floating').

Assuming the test LED is correctly off with the test probe in point number 3, type: `SETPIN(2), OFF` (Enter), `SETPIN(3),DOUT` (Enter), `PIN(3)=1` (Enter). This should turn on the test LED once again.

You then need to work through the above process that is summarised as:

- Probe into point `x`
- `SETPIN(x),DOUT`
- `PIN(x)=1`
- Check LED on
- Move probe to next point `y`
- Check LED off
- `SETPIN(x),OFF`
- `SETPIN(y),DOUT`
- `PIN(y)=1`
- check LED on

Replace `x` and `y` above with the relevant pin numbers; so `x` would become 3, and `y` becomes 4. Hopefully this is fairly straightforward, but remember the pins are not always the next value up, so when checking point number 7 (for `x`), `y` will be 9 (and not 8). Refer to J12 in Fig.2 for the numbering sequence you need to use. When you finally reach 26 (for `x`), the LED should turn on when you type `PIN(26)=1`. If it does, then testing is complete.

Configuring your *MKC* and TeraTerm

To make things more user-friendly, we will tweak some settings in both the Micromite, and in TeraTerm (refer to the screen shots in Fig.3). Begin by connecting your *MKC* to TeraTerm, (remember to select the correct COM Port number, and also ensure baud speed is set to 38,400) and then press the Reset button on the *DM* (top-left button). You should see the Micromite and copyright message, as in the two lines shown in *Part 3* (Fig.9 on page 65). This means that you are successfully connected to the PC. Now press the 'C' key while holding down Ctrl and you should see the command prompt (>).

Next, type the following (strictly in the sequence shown). First, type `NEW` (Enter). This will remove the program in your *MKC*. If you had the *MKC* test software installed, then on the next power-up there will be no program to automatically run, hence you will just see the screen, as shown in *Part 3*, Fig.9.

Next, type in: `OPTION DISPLAY 40,125` (Enter). This will give a bigger screen area to work with. And finally, type: `OPTION BAUDRATE 115200` (Enter). Be careful to check that you type exactly '115200'. This setting will speed up the console link between the *MKC* and TeraTerm (think of it as improving the screen refresh rate in TeraTerm). Note that when you next press Enter you will appear to lose connection to TeraTerm – don't panic! We will shortly change the baud rate in TeraTerm to match.

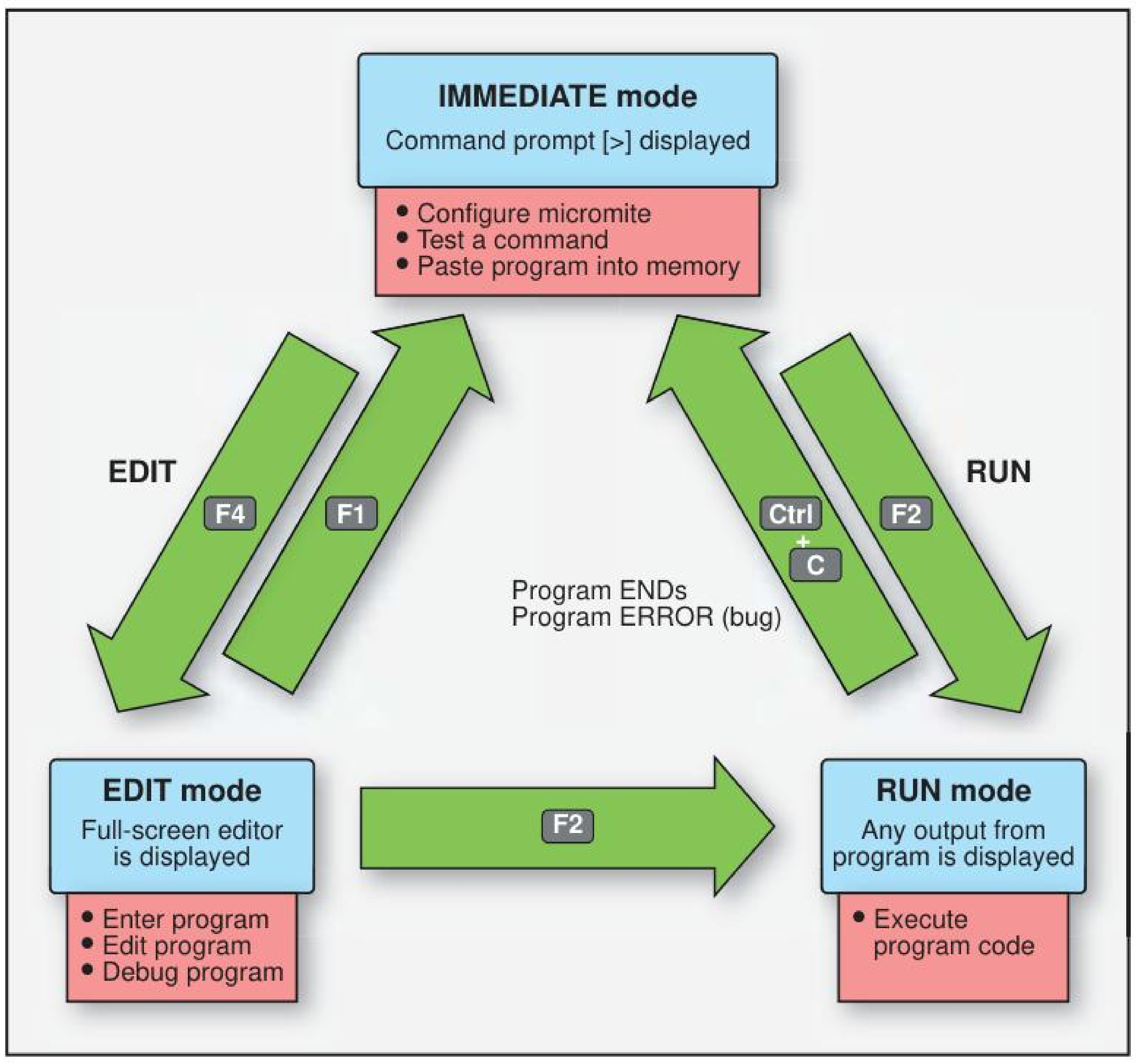


Fig.4. Summary of the three Micromite modes and how to switch between them.

In TeraTerm, select the Setup tab and work down the list making sure your settings match those as shown in the screen shots in Fig.3. Finally, it is important you select 'Save setup'; and if you followed everything correctly then the next time you start TeraTerm, you will see a nice blue screen and you should be communicating once again with your *MKC* but at the faster baud rate of 115200. Press the reset button on the *DM* and ensure you see the welcome message and the cursor (>).

The basics of MMBASIC

We have previously recommended that you download the excellent *Getting Started with the Micromite* guide, and also the *Micromite User Manual*. If you have not yet done so, then please go ahead and download them from **geoffg.net/micromite.html** (scroll to bottom of page and they are the first couple of items under the 'Downloads' heading).

It's not possible in the limited space here to cover everything about coding, so we will be taking things easy and will learn commands and techniques along the way. The idea is to show working examples with the use of some basic (cheap) components. For those of you interested in exploring topics at a greater level, then please refer to the manuals – they contain a lot of information.

That said, the following are some essential concepts that we need to understand before moving on.

- An electronics project/gadget can generally be broken down into three elements: Input – Process – Output. The *MKC* is the Process element and adds intelligence.
- The intelligence is in the form of a program. The Micromite only ever contains a single program – there is no file system with multiple programs stored in memory
- Flash memory is used to store this single program, so when powering up the Micromite, the program is already in memory and immediately available
- A program comprises lines of code. A line of code contains instructions (commands) that the Micromite executes. Commands can be typed in upper or lower case; however, it is generally easier to read code if it is typed in upper case
- When a program is 'run', it will start at the first line of code and will generally flow down to the next line of code. However, some commands can cause the program to 'jump', 'branch', or 'loop' to other areas of the program depending upon a condition or test
- With MMBASIC, there are many ways to achieve the same end result. No one way is right or wrong; but some ways may be better than others
- Some commands have an immediate visible output (such as turning on an LED), while others are 'configuration'; ie, not a visible output, (such as setting an I/O pin to a digital output type)

One of the most important things to understand about the Micromite is

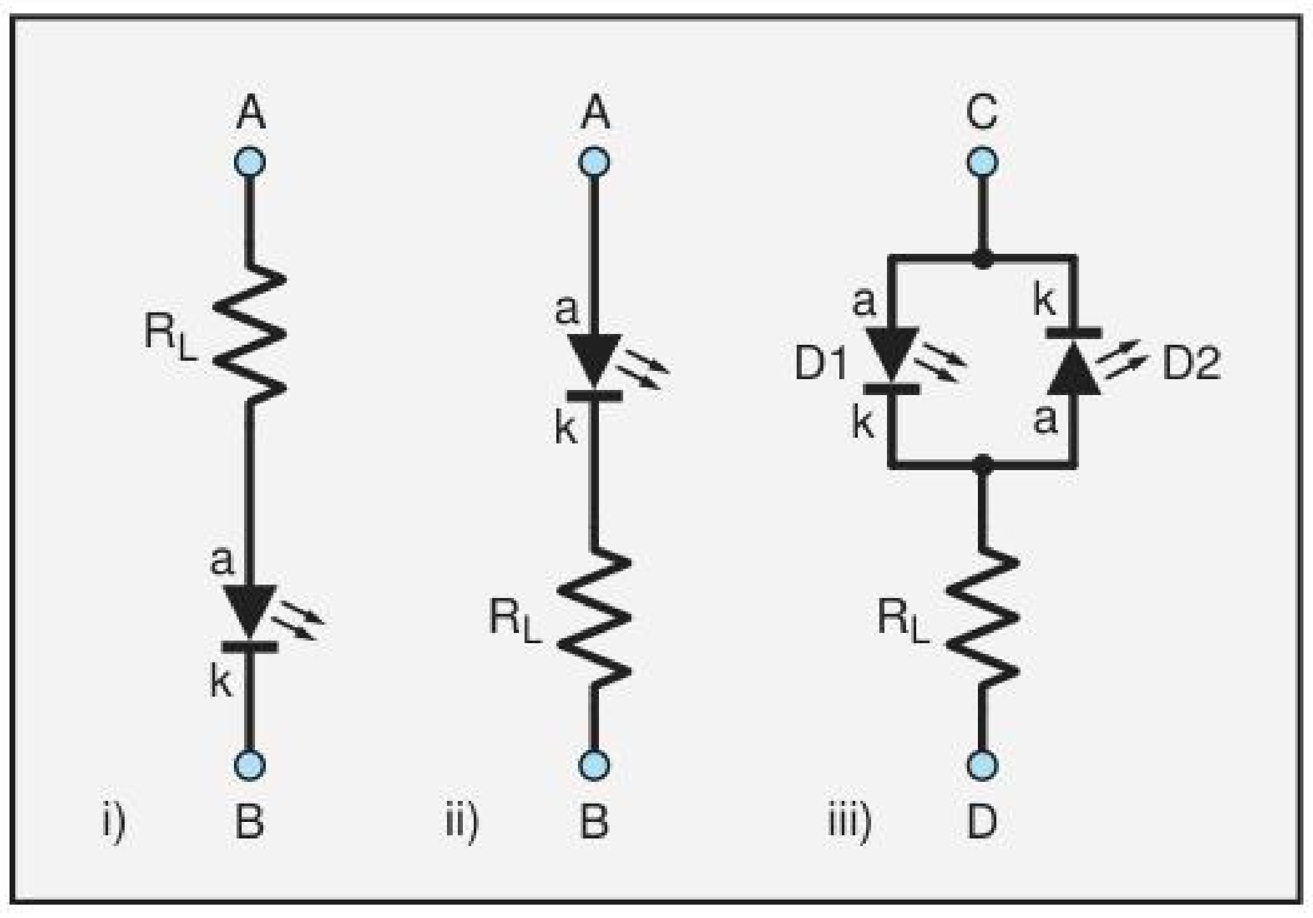


Fig.5. Various ways used to connect LEDs.

that at any time, it is only ever in one of three modes. It is either waiting for an instruction from you (command prompt displayed), or, you are entering a program (full screen editor displayed), or, it is running the program in memory. These three modes are referred to as IMMEDIATE mode, EDIT mode, and RUN mode, as shown in Fig.4.

To switch between modes, you either type a single command (if at the command prompt), or you press a function key (F1, F2, or F4), or Ctrl and C. There are other key-press combinations based around the Ctrl key for those of you with keyboards lacking function keys (refer to *User Manual – Full Screen Editor* section).

One important point mentioned in *Part 1* – the Micromite makes it extremely quick and easy to change a program (in the Editor), and then immediately run the program to see the result of the change. Should you need to make further changes (you may be adjusting the speed of a flashing LED which is controlled by a single parameter) then pressing the single relevant function key switches modes immediately and you can make the change as needed. Most other microcontrollers require you to make the change on a PC, then compile your code, then upload your code to the microcontroller, then run the code. Any further changes require this whole sequence to be repeated (which can take a considerable time).

Please take a moment to study Fig.4 carefully, and then for a little practice performing switching modes, turn to the bottom of page 11 in the *User Manual* (*A Simple Program* section) and create, run, and edit the one-line 'Hello World' program shown.

Inputs and outputs

We will soon move on to controlling and sensing basic digital inputs and outputs, but before that, let's first examine an input and an output that involves no components connected to the I/O pins. I am talking about the console; ie, TeraTerm. The screen is an output, and the keyboard is an input.

To get something displayed on the TeraTerm screen we can use the command `PRINT`. Go ahead and try this. At the command prompt type: `PRINT 22/7` (Enter). You should see `3.14286` displayed on the TeraTerm screen (along with the cursor on the next line waiting for the next instruction from you). In the Immediate mode, when you press the Enter key, anything you type on the line will be acted upon immediately. Press some random keys and then Enter and see what happens. You should see an error message along with a clue as to what the error is.

The TeraTerm keyboard can be used as an input with the command `INPUT`. To quickly demonstrate, type the following as a single line at the command prompt:

```
INPUT x : PRINT "Number entered was:"; x
```

When you press Enter you will see a question mark appear. This is MMBASIC waiting for you to input a number. Type in a number and then press the Enter key. You will then see an 'output' displayed (because of the `PRINT` command). This is a very simple example of how to use the TeraTerm console as an input (via the keyboard), and as an output (via the screen).

Digital outputs

A digital output device is something that is either 'on' or 'off'. An obvious component to demonstrate this is an LED; however, it could just as easily be a motor, a buzzer, a relay… Note that the device being controlled may well need some interface electronics between it and the Micromite. For example, the Micromite has not got the power to turn on a powerful mains motor, but this could be done with a relay controlled by the Micromite (and the mains motor connected to the relay contacts).

Back to controlling an LED; please now refer to Fig.5 i). To turn the LED on, point A is connected to +V, and point B to 0V. The resistor is required to limit the current flowing through the LED (as is standard). However, to have the Micromite control the LED we connect point A to an I/O pin (instead of +V), and connect point B to 0V. This is exactly how the LED test circuit works (as used when I/O pins on the *MKC* and *DM* were tested).

Go ahead and connect a resistor and an LED to any one of the 19 available I/O pins (refer to Fig.2 for pin number/locations). It doesn't matter if you switch round the positions of the resistor and the LED (as in Fig.5 (ii)) – the same logic still applies.

There are two commands we need to use – one will configure the relevant pin to be a digital output, and the other command will turn the pin on (+V), or off (0V).

At the command prompt, type `SETPIN(2),DOUT`, substituting the value 2 with whatever pin number you are using. When you press Enter, nothing will appear to happen; however, you have now configured the pin for use as a digital output (`DOUT`).

Now type `PIN(2)=1` again substituting the value 2 with the relevant pin number. This time you should see something happen – the LED will turn on. If not, check connections and commands to ensure you have used the correct positions and pin number value.

Now type `PIN(2)=0` (using your pin number) and the LED will turn off. That's it, you now have software control of hardware via a digital output pin.

To make it more interesting, we will now enter a small program into the Editor and add a bit of intelligence. Note that while entering a line of code (into the Editor), when you press Enter the command is not acted upon as it would be at the command prompt. The Editor is just a place to enter a program; and things will only happen when you *run* your program.

At the command prompt, type `NEW` (to delete any code currently in the Micromite) and then type `EDIT` (or press F4). You should now have a blank screen with just a line of words at the bottom. Go ahead and type the following (pressing Enter to move to the next line down):

```
SETPIN(2),DOUT   ' configure pin 2 as digital output
PIN(2)=1         ' switch pin 2 ON
PAUSE 500        ' delay for 500mS (1/2 second)
PIN(2)=0         ' switch pin 2 OFF
PAUSE 500        ' 0.5 second delay
```

Note that the command `PAUSE` just adds a delay. The Micromite will sit there doing nothing for the time specified in milliseconds. There are 1000 milliseconds (ms) in one second; so 500ms is half a second.

An important aspect of the Editor is that comments can be added to a program by using the '`'`' character. Anything on a line of code typed after the '`'`' is ignored by the Micromite.

Now move your LED over to I/O pin 2 and press F2 to automatically save and run your program. If everything works correctly, then you will see the LED turn on for half a second, and then turn off. Because the program was able to continue and reach the end of the code, the Micromite automatically returned back to the command prompt. However, if there was an error of some kind, then a warning will be displayed on the TeraTerm screen. If this happens, press F4 and the cursor will be placed at the start of the offending line of code. Check you entered things exactly as shown above, correcting any errors. Then press F2 to run the program again.

To make things a bit more interesting we will now create what is called a 'loop', allowing the LED to continually switch on, then off. Put another way, it will make the LED continually flash. To do this we use the command pair: `DO` and `LOOP`. `DO` needs to be added after the `SETPIN` line of code and marks the start of the loop; and `LOOP` goes at the end of the program so that the program jumps back to the line of code immediately after the `DO` command.

Go to the Editor (F4) and position the cursor at the end of the first line of code. Note: you can use the End key to position the cursor at the end of the current line of code; and a second press of the End key will position the cursor at the end of the very last line of code. Likewise, with the Home key; one press to go to the start of the current line, and a second press to go to the start of the first line of code. Alternatively, use the four arrow keys to move the cursor. You cannot use the mouse to scroll, or to position the cursor – this is a limitation of TeraTerm.

So, with the cursor at the end of the first line of code (the line containing the `SETPIN` command), press Enter and a new blank line is inserted. Now type `DO` and then press the End key twice to go to the end of your program and press Enter, and then type `LOOP`. Your program should now look something similar to that shown in the *User Manual* on page 12 (but with a different pin number, and with different `PAUSE` values). It is worth indenting the lines of code that are between `DO` and `LOOP` (as shown in the program listing on page 12). You can use the tab key to do this, or use the spacebar. Indenting makes it much easier to read the code (and has zero impact on performance because the Micromite just ignores spaces at the start of a line).

Press F2 and this should result in the LED turning on for half a second, then off for half a second, and then continuing to flash like this until you either remove power, reset the Micromite, or press Ctrl-C.

Experiment with changing the values of the two `PAUSE` commands to vary the flash rate (they don't need to be the same as each other). This is also good practice for switching between the three Micromite modes, while seeing the effects of changing values in your code.

We recommend you read the rest of page 12 (*Setting the AUTORUN option*). In summary, at the command prompt, type `OPTION AUTORUN ON` and then reset the Micromite with the reset button on the *DM*. You will see the start-up message, and then the program will automatically run (LED flashes). Ctrl-C will stop the program running and return you to the command prompt. `OPTION AUTORUN OFF` will prevent the program in the Micromite from running automatically at power up.

Referring back to Fig.5 i) and ii), point A is controlled by the command `PIN(x)=1` (or `0`) to turn the LED on (or off) – with `x` replaced by the relevant pin number. Now consider connecting point B to an I/O pin (instead of connecting it to 0V). Let's connect it to pin y (ie, not the same pin as point A). You may ask why do this, but bear with us. The LED will only come on when `PIN(x)=1` and `PIN(y)=0`. If `PIN(y)=1`, then the LED is off, irrespective of `PIN(x)` setting. If this sounds complex, just liken it to point B being connected to +V. You will not be able to turn the LED on while point B is at +V.

The reason for explaining this is that we can do a neat trick if we connect two LEDs back to back (to a single resistor) as shown in Fig.5 iii). Assume that point C is connected to I/O pin `x`, and point D is connect to I/O pin y. Now we have four combinations of outcome depending on how the Micromite sets `PIN(x)` and `PIN(y)`:

`PIN(x)=0` and `PIN(y)=0` – both LEDs off
`PIN(x)=1` and `PIN(y)=0` – LED D1 on, D2 off
`PIN(x)=0` and `PIN(y)=1` – LED D2 on, D1 off
`PIN(x)=1` and `PIN(y)=1` – both LEDs off

At first glance this may not seem very useful, but now consider what happens if we rapidly switch between `PIN(x)=1` with `PIN(y)=0` and then switch to `PIN(x)=0` with `PIN(y)=1`. It will result in both LEDs being on; albeit they will be slightly dimmer in appearance due to the rapid switching between on and off (but our persistence of vision will make it appear as if both LEDs are on).

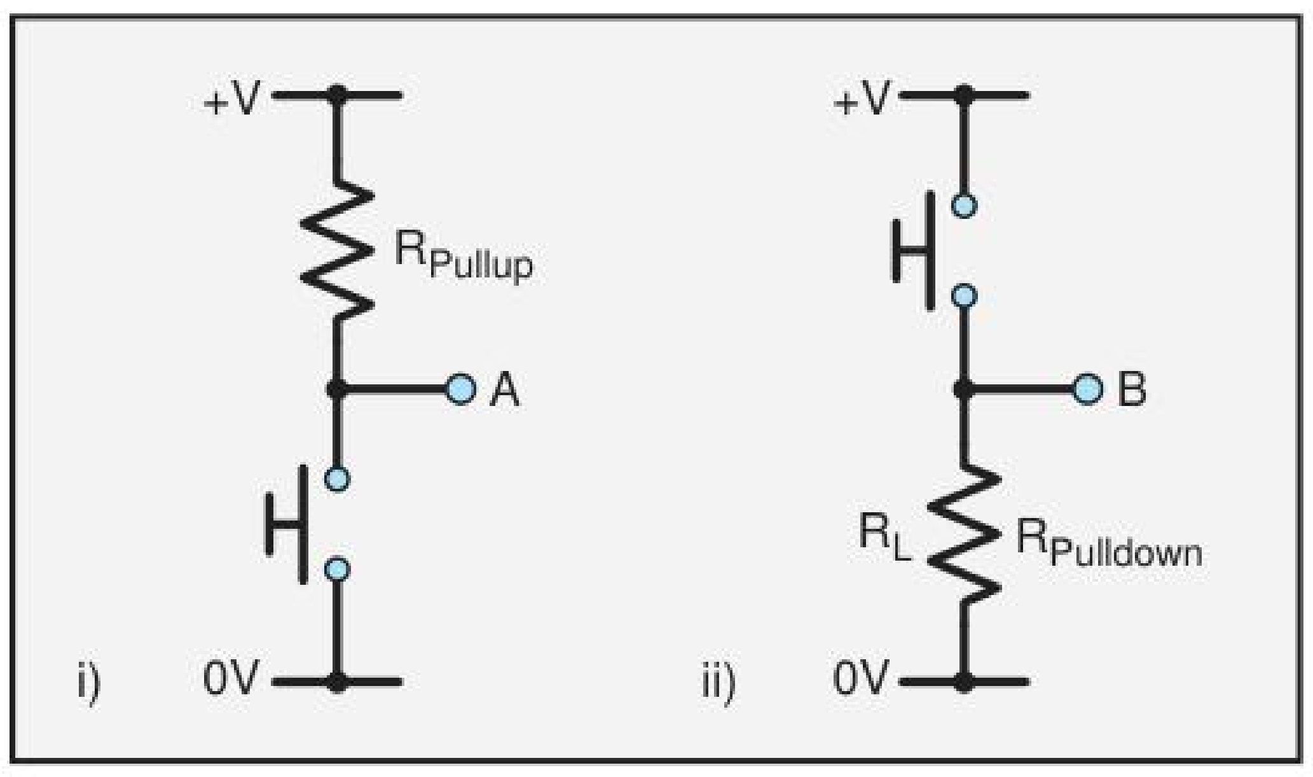


Fig.6. Two different ways to connect a button to a digital input pin: i) active low, ii) active high.

So with just two I/O pins, and one resistor, we can alter our code to result in any combination of LEDs being turned on and/or off.

Don't worry if this seems confusing; it is mentioned here because it is a technique that will be used in next month's *Electronic Dice* program, making it possible to control seven LEDs with just three I/O pins (and only four resistors).

Digital inputs

A digital input pin allows MMBASIC to be able to monitor a device that is in one of only two states. A switch or button is a great example of a digital input device – it can be either open or closed. Referring to Fig.6 i), this is a typical connection of how we can detect if a button is being pushed or not. When the button is not being pressed, point A will be at +V (regarded as a high logic level) due to the pull-*up* resistor. However, when the button is pressed, point A is then directly connected to 0V, which is regarded as a low logic level (point A is 'active low'). A resistor value in the region of 10kΩ will ensure that when the button is pressed, then only a small amount of current flows between +V and 0V rather than shorting it out.

Fig.6 ii) is similar to i), but the logic levels are now working in reverse. Point B is at a low logic level when the button is not being pressed (thanks to the pull-*down* resistor), and at a high logic level when the button is pressed (point B is 'active high').

To demonstrate the above, build the circuit shown in Fig.6 i) and use it to control the flashing LED program from the previous section. You need a button or switch, and a resistor with a value between 470Ω and 100kΩ. If you don't have a button or switch, then simply use two jumper wires and you can touch them together to simulate a button press. Use a breadboard, a small piece of stripboard, or even just the leads of the resistor to assemble the circuit shown in Fig.6 i). Connect the +V side of the pull-up resistor to the middle 4-way socket row, J8 (this is at a 3V3 level). Connect one side of the button to 0V (J7, the left hand 4-way socket row), and point A to I/O pin 14.

Now insert the following three lines of code to the very beginning of your program:

```
SETPIN(14),DIN ' make I/O pin 14 a digital input
waiting:       ' this is a label we are defining
IF PIN(14) = 1 THEN GOTO waiting ' test High (1)
```

Before you press F2 to run this, there are three new things introduced here that briefly need explaining. The `SETPIN` command is configuring the I/O pin 14 as a digital input with the use of `DIN` (as opposed to `DOUT`). More information is available in the *User Manual* on page 39 (*Using the I/O pins*).

A label called `waiting:` has been defined and simply gives us a point in our code that can be referenced should we wish to jump to that part of the code. Remember, program flow normally goes from top to bottom, but we can jump to other parts should we need to. Labels give us this ability and need to be a word that is not the same as an MMBASIC command. It also needs to have a colon character immediately after the last character. Finally, the `IF...THEN...` is an example of a 'test condition'. What this is doing is quite straightforward. It is testing the state of

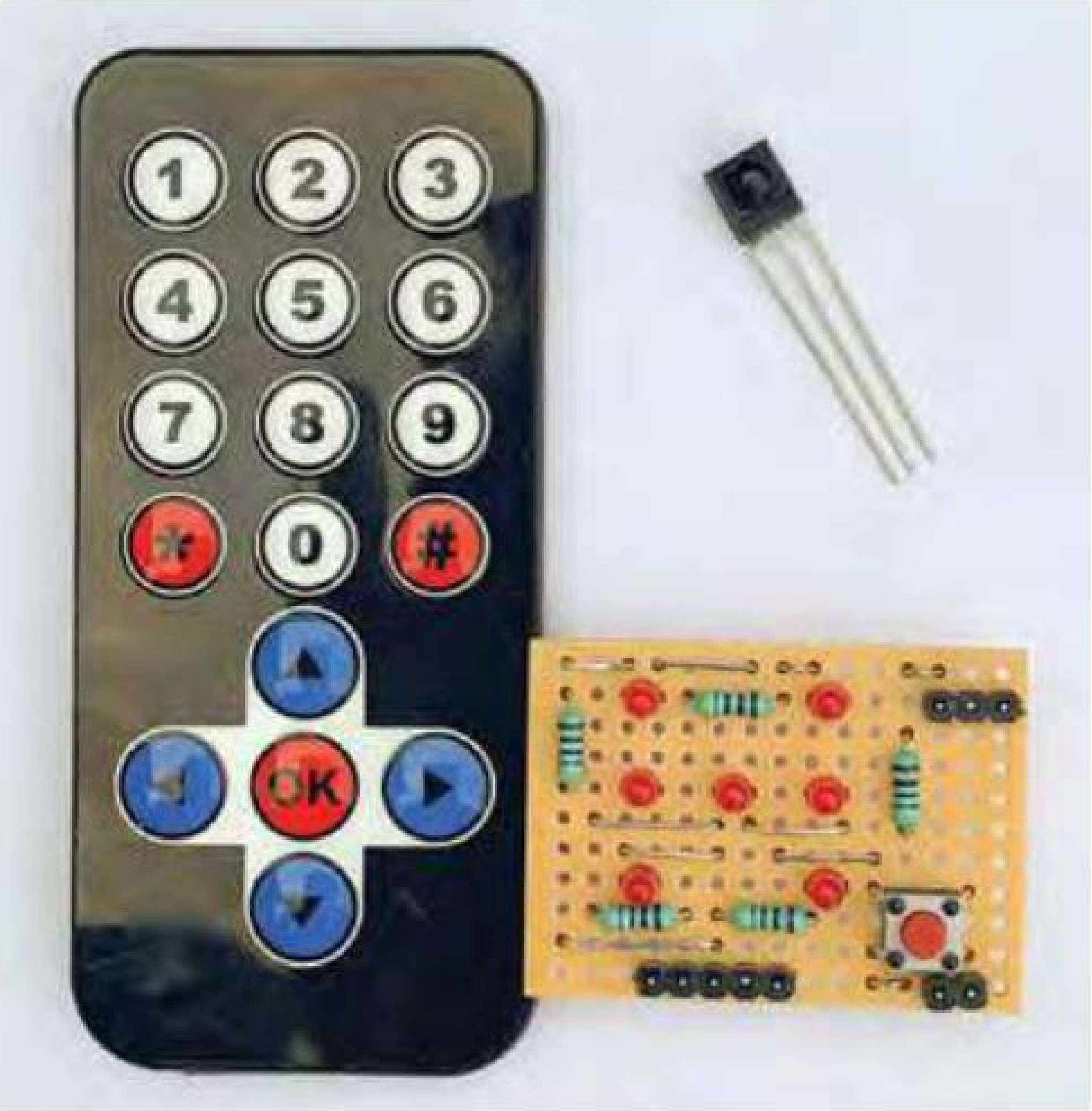


Fig.7. A complete kit of parts for the *Electronic Dice*. The kit includes a mini-IR controller and the relevant IR receiver which will be used in many of our mini-projects. It is available from micromite.org

I/O pin 14, and if it is at a high logic level (which is represented by the value 1), then the program flow will `GOTO` the label `waiting:` – otherwise, it will continue on down to next line of code. For our circuit, this means while the button is not pressed the program flow will continually jump to `waiting:` The only way the program will go past this `IF...THEN...` line of code is when `PIN(14)=0`, ie, the button is pressed.

So go ahead and press F2 to run your new code. At first, nothing will appear to happen because the LED is not flashing. However, the program flow is following your instructions precisely and it is only when you press the button that the LED will begin to flash.

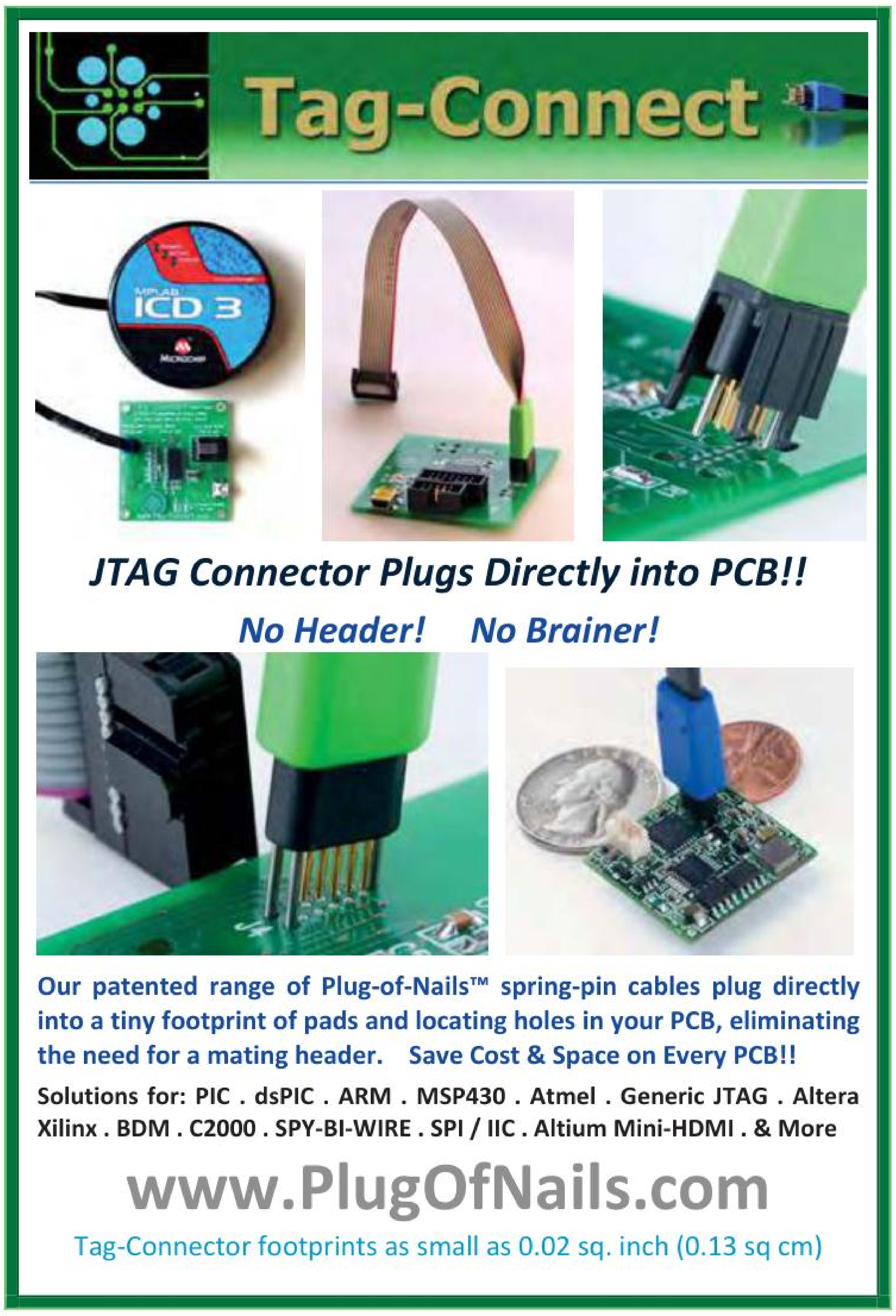
We will now use a feature that demonstrates how you can remove a hardware component simply by modifying the `SETPIN` command. Refer to Fig.6 i), which is the circuit you have currently working. Thanks to MMBASIC we can now physically remove the pull-up resistor from our circuit and use an additional parameter with the `SETPIN` command; as follows: `SETPIN(14),DIN,PULLUP`

Run your code once again and check it all works (the LED only starts flashing once you press the button).

Having worked through these simple digital input and output building-block examples, you can begin to see just how easy it is to use a Micromite to add intelligence to a project. Furthermore, by just changing your code slightly (and leaving the hardware the same), you can add features that are simply not possible with hard wiring alone.

Next month

In the next part we will introduce a few more new commands and techniques before building the *Electronic Dice* mini-project. We will show you how easy it is to add an infrared remote keypad to a Micromite, and will use this to make a simple dice game. In the meantime, please enjoy experimenting!



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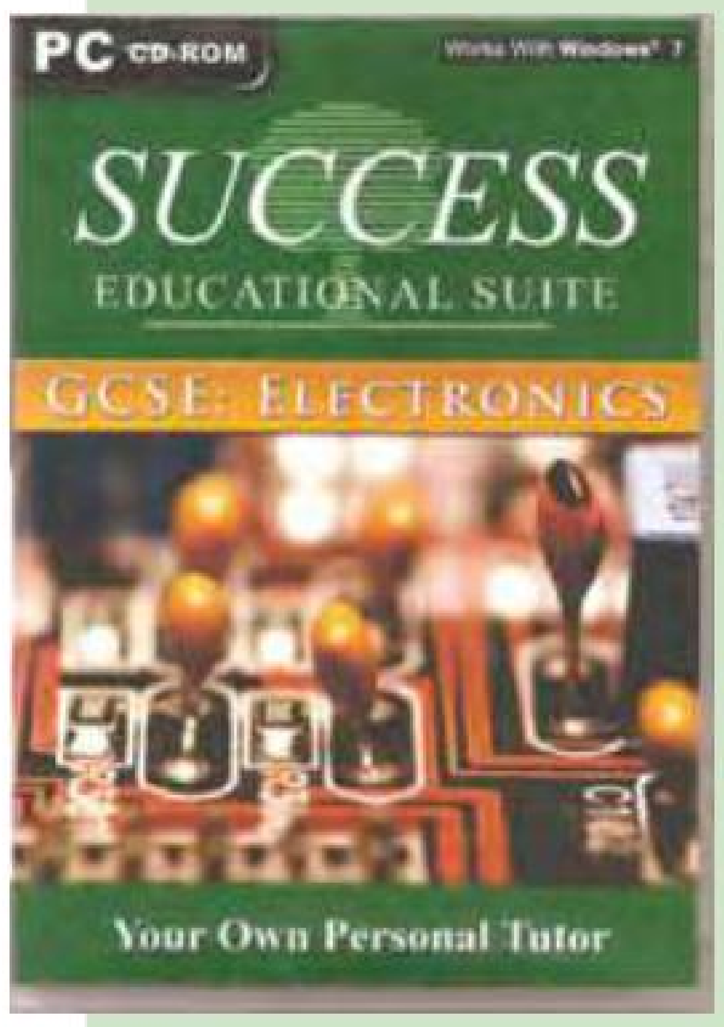

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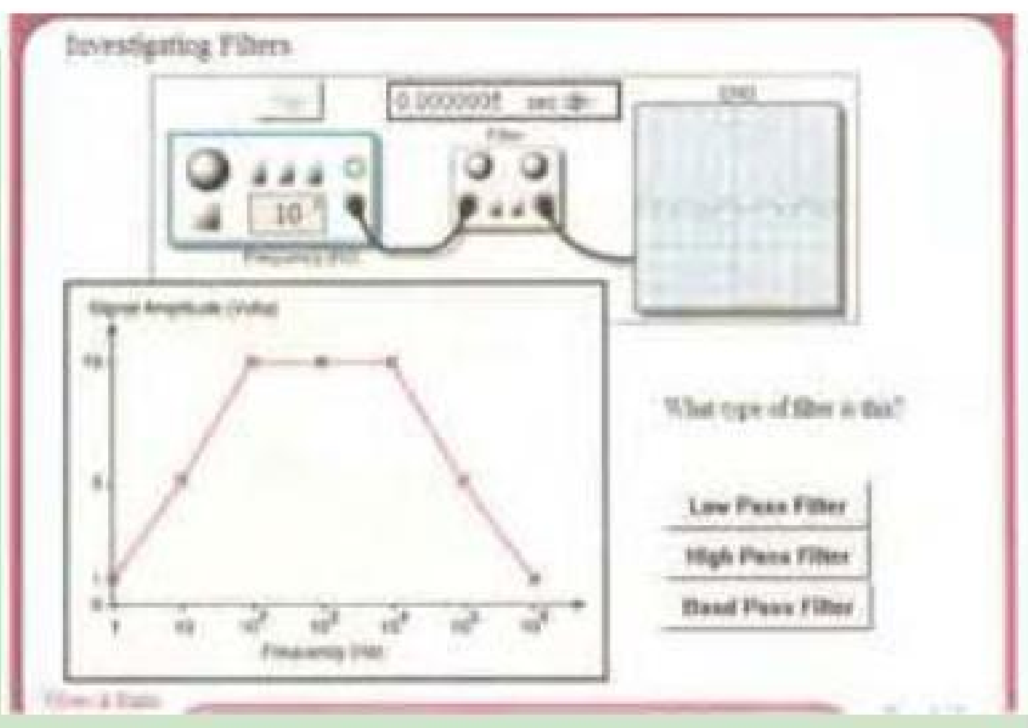

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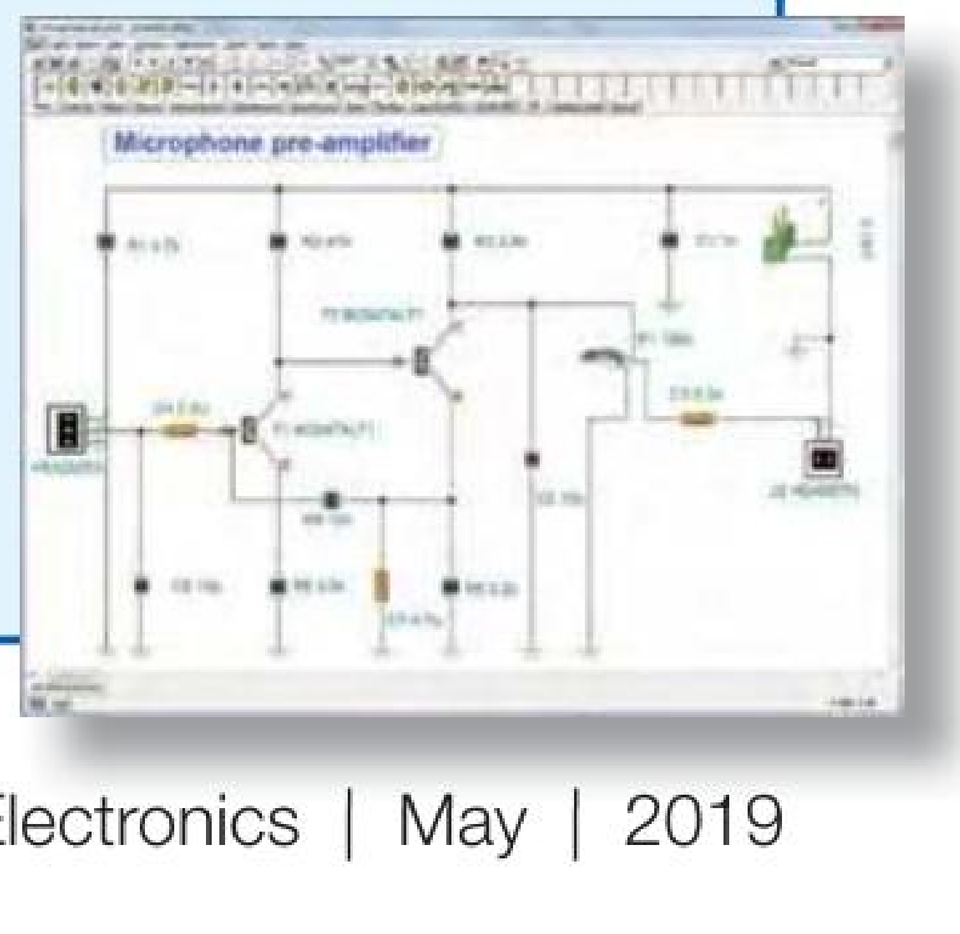

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(Formerly PICtutor)

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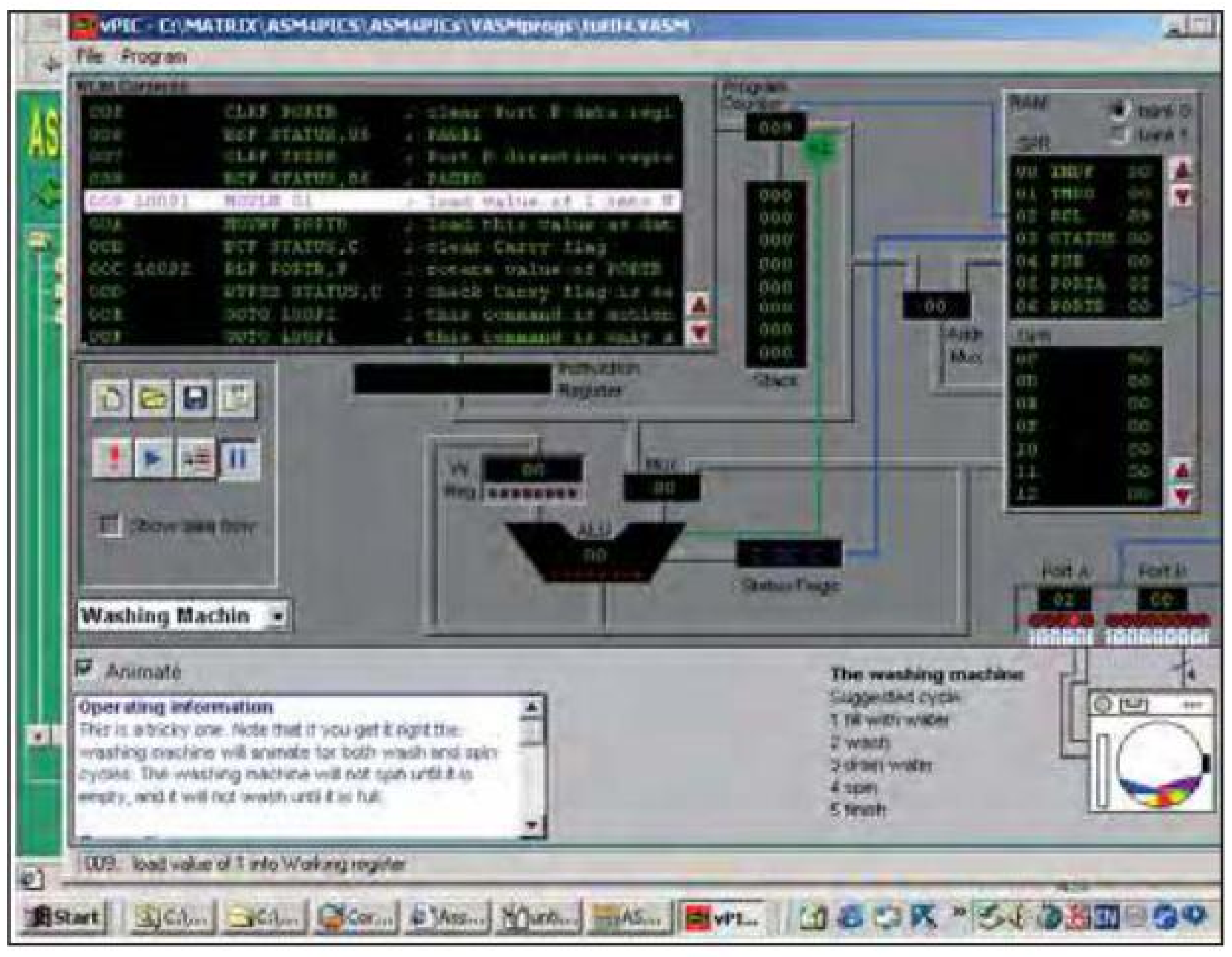

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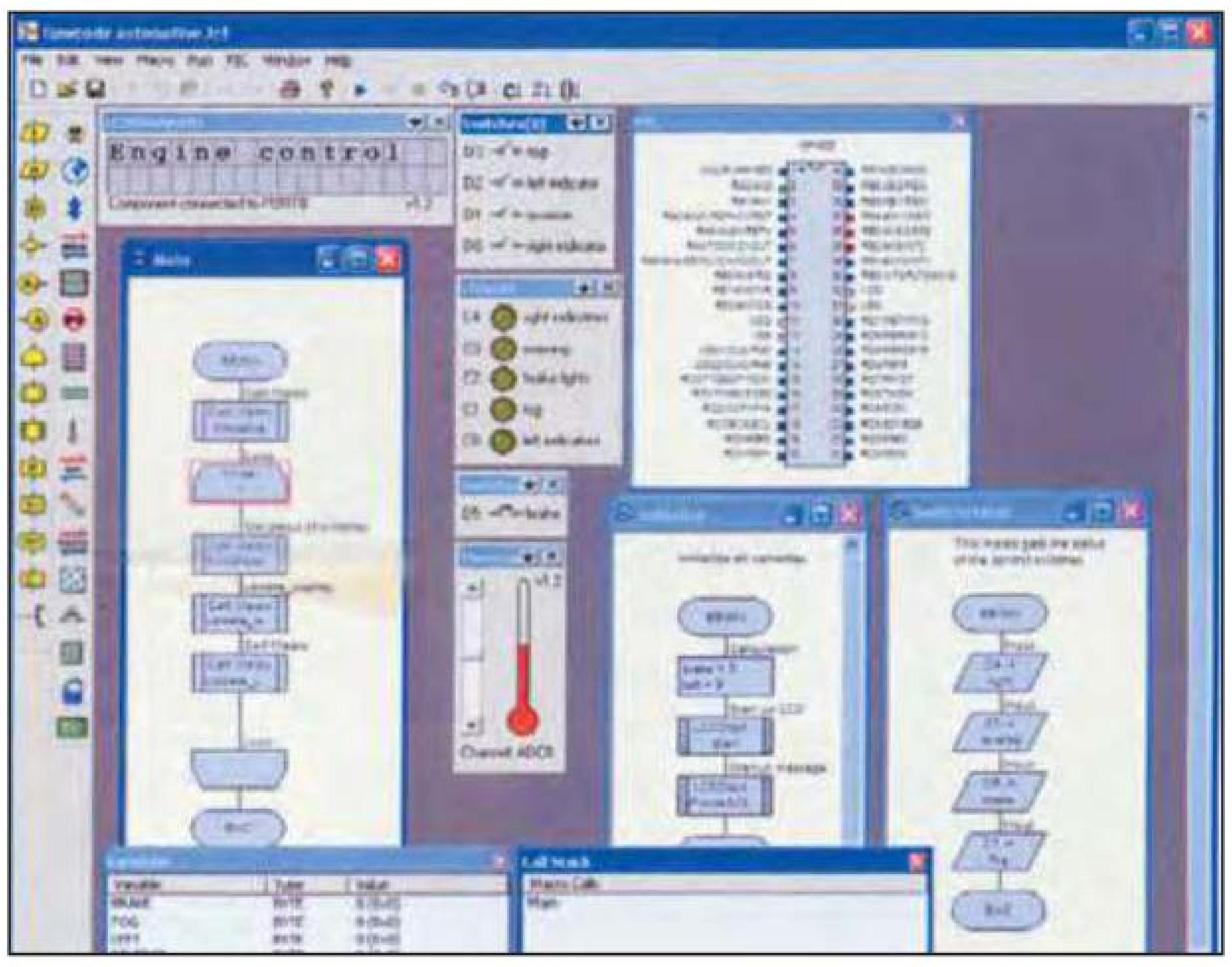

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This *Teach-In* series of articles was originally published in *EPE* in 2008 and, following demand from readers, has been collected together in the *Electronics Teach-In 2* CD-ROM.

The series is aimed at those using PIC microcontrollers for the first time. Each part of the series includes breadboard layouts to aid understanding and a simple programmer project is provided.

Also included are 29 *PIC N' Mix* articles, also republished from *EPE*. These provide a host of practical programming and interfacing information, mainly for those that have already got to grips with using PIC microcontrollers. An extra four part beginners guide to using the C programing language for PIC microcontrollers is also included.

The CD-ROM also contains all of the software for the *Teach-In 2* series and *PIC N' Mix* articles, plus a range of items from Microchip – the manufacturers of the PIC microcontrollers. The material has been compiled by Wimborne Publishing Ltd. with the assistance of Microchip Technology Inc.

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For Beginners of Any Age
Python Coding on the BBC micro:bit
Jim Gatenby

PYTHON CODING ON THE BBC MICRO:BIT

Jim Gatenby

Python is the leading programming language, easy to learn and widely used by professional programmers. This book uses MicroPython, a version of Python adapted for the BBC Micro:bit.

Among the many topics covered are: main features of the BBC micro:bit including a simulation in a web browser screen; various levels of programming languages; Mu Editor for writing, saving and retrieving programs, with sample programs and practice exercises; REPL, an interactive program for quickly testing lines of code; scrolling messages, creating and animating images on the micro:bit's LEDs; playing and creating music, sounds and synthesized speech; using the on-board accelerometer to detect movement of the micro:bit on three axes; glossary of computing terms.

This book is written using plain English, avoids technical jargon wherever possible and covers many of the coding instructions and methods which are common to most programming languages. It should be helpful to beginners of any age, whether planning a career in computing or writing code as an enjoyable hobby.

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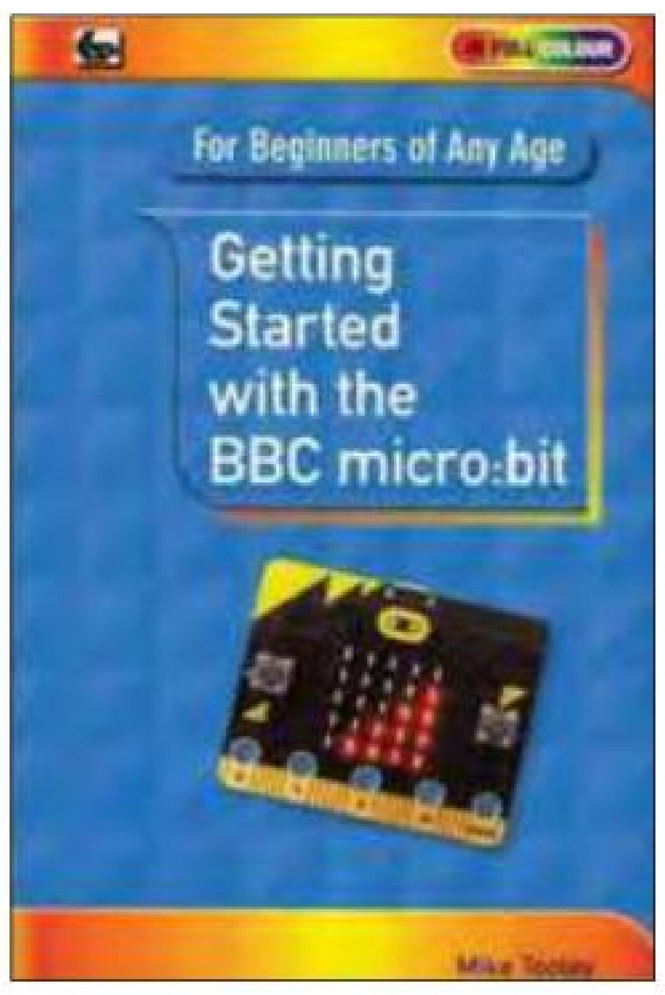

GETTING STARTED WITH THE BBC MICRO:BIT

Mike Tooley

Not just an educational resource for teaching youngsters coding, the BBC micro:bit is a tiny low cost, low-profile ARM-based single-board computer. The board measures 43mm × 52mm but despite its diminutive footprint it has all the features of a fully fledged microcontroller together with a simple LED matrix display, two buttons, an accelerometer and a magnetometer.

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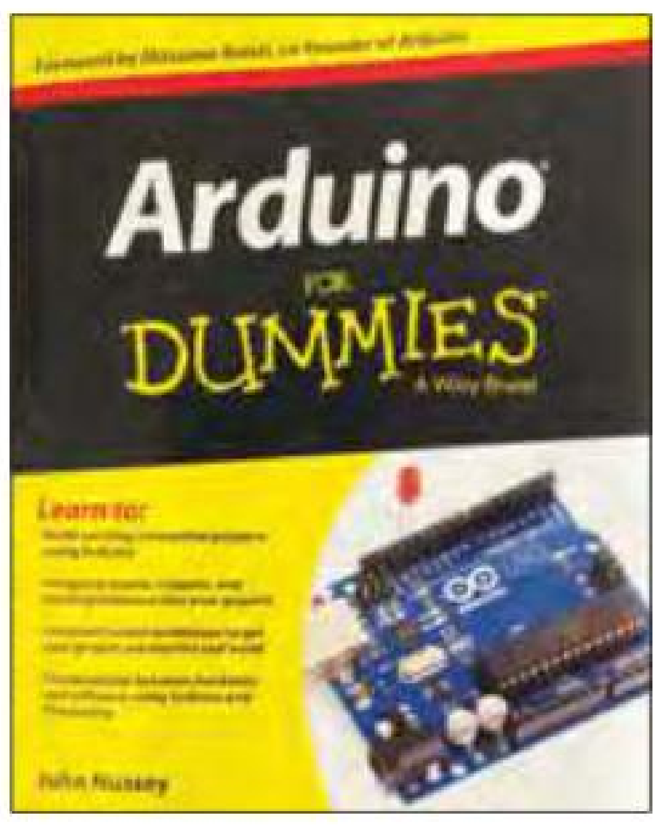
ARDUINO

Teach-In 2016

See opposite for our popular introduction to the Arduino



ARDUINO FOR DUMMIES
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Arduino is no ordinary circuit board. Whether you're an artist, a designer, a programmer, or a hobbyist, Arduino lets you learn about and play with electronics. You'll discover how to build a variety of circuits that can sense or control real-world objects, prototype your own product, and even create interactive artwork. This handy guide is exactly what you need to build your own Arduino project – what you make is up to you!

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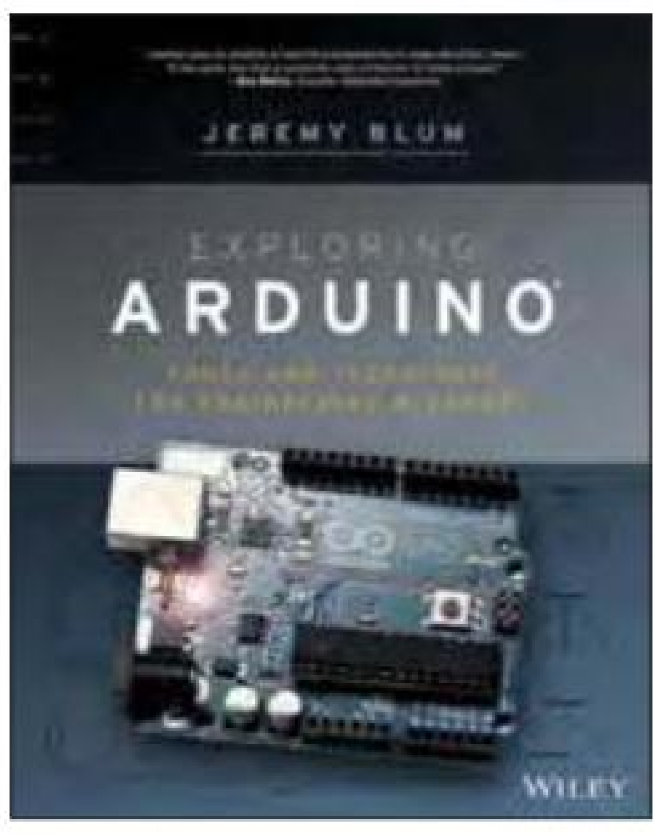

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Jeremy Blum

Arduino can take you anywhere. This book is the roadmap.

Exploring Arduino shows how to use the world's most popular microcontroller to create cool, practical, artistic and educational projects. Through lessons in electrical engineering, programming and human-computer interaction, this book walks you through specific, increasingly complex projects, all the while providing best practices that you can apply to your own projects once you've mastered these. You'll acquire valuable skills – and have a whole lot of fun.

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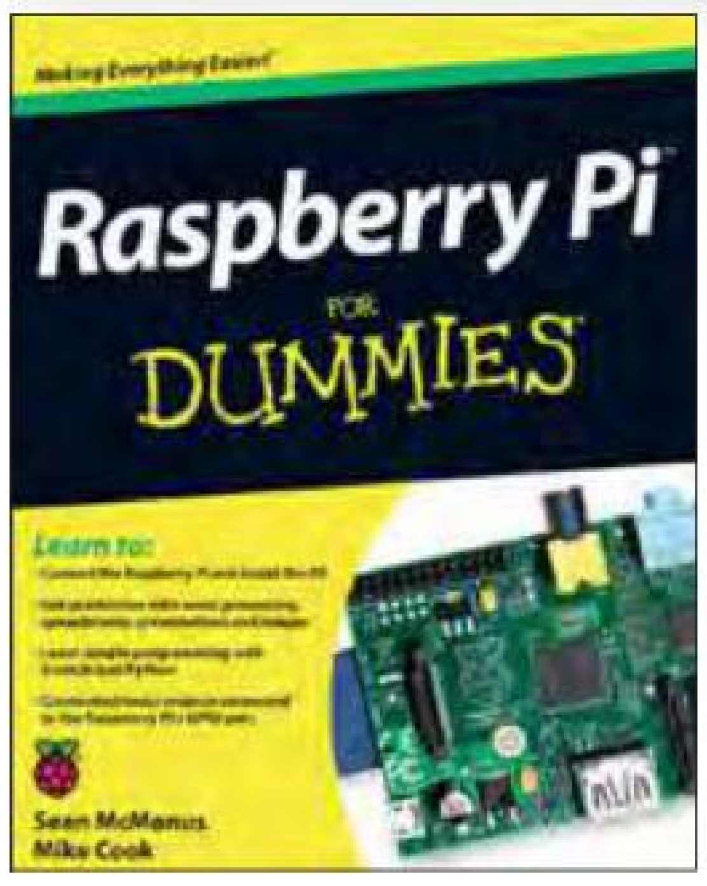

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Sean McManus and Mike Cook

Write games, compose and play music, even explore electronics – it's easy as Pi! The Raspberry Pi offers a plateful of opportunities, and this great resource guides you step-by-step, from downloading, copying, and installing the software to learning about Linux and finding cool new programs for work, photo editing, and music. You'll discover how to write your own Raspberry Pi programs, create fun games, and much more!

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Mike & Richard Tooley

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The CD-ROM also contains all the necessary software for the series so that readers can get started quickly and easily with the projects and ideas covered.

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The Arduino offers a remarkably effective platform for developing a huge variety of projects; from operating a set of Christmas tree lights to remotely controlling a robotic vehicle through wireless or the Internet. Teach-In 8 is based around a series of practical projects with plenty of information to customise each project.

This book also includes *PIC n' Mix*: PICs and the PICkit 3 - A Beginners guide by Mike O'Keefe and *Circuit Surgery* by Ian Bell - State Machines part 1 and 2.

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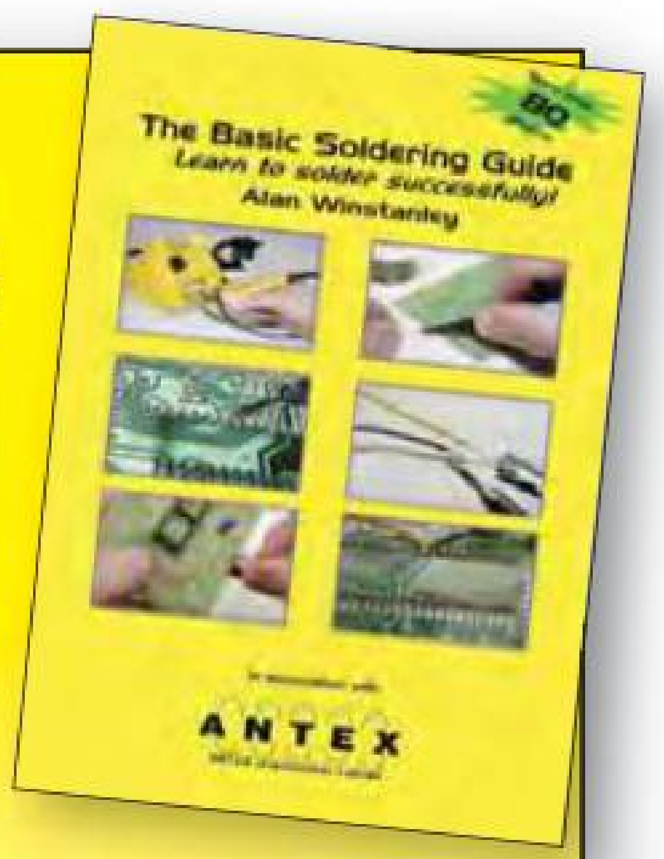

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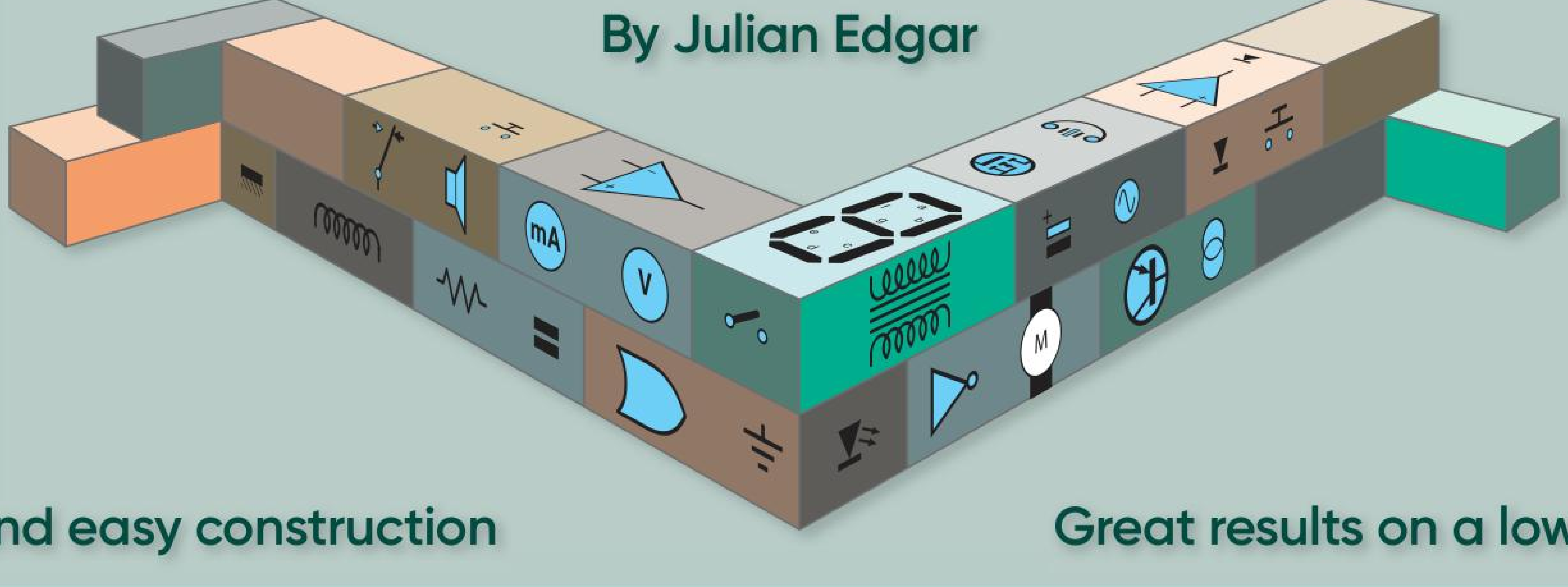
Electronic Building Blocks

By Julian Edgar



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Multi-pulser

The item discussed this month is a multi-purpose variable duty cycle and frequency driver with 5A capability that can perform lots of different functions. Examples? It can be used to help teach electronics; test pulse-width modulated (PWM) solenoids and car fuel injectors; act as a DC motor speed control or light dimmer; or even a fully-configurable DC light flasher.

We'll look at each of those uses in a moment, but first, what do we actually have here? In fact, it's two modules: a frequency generator module, and a MOSFET power board – together, they form the 'Multi-pulser'!

Frequency generator module

The frequency generator module is a self-contained device that outputs a variable duty cycle, variable frequency, square wave signal. Control of frequency (1Hz – 150kHz) and duty cycle (0-100%) is via front-mounted up/down pushbuttons, and the selected frequency and duty cycle are shown on a back-lit LCD. Power supply can be anything from 3.3 – 30V. The selected frequency and duty cycle settings are saved, even with power switched off. Maximum output power of the module is 30mA, and the PWM output swing reflects the supply voltage. The module is designed to be flush-mounted on a panel, requiring a cut-out of 72 x 39mm.

FREQ
DUTY
PWM FREQUENCY METER XY-PWM

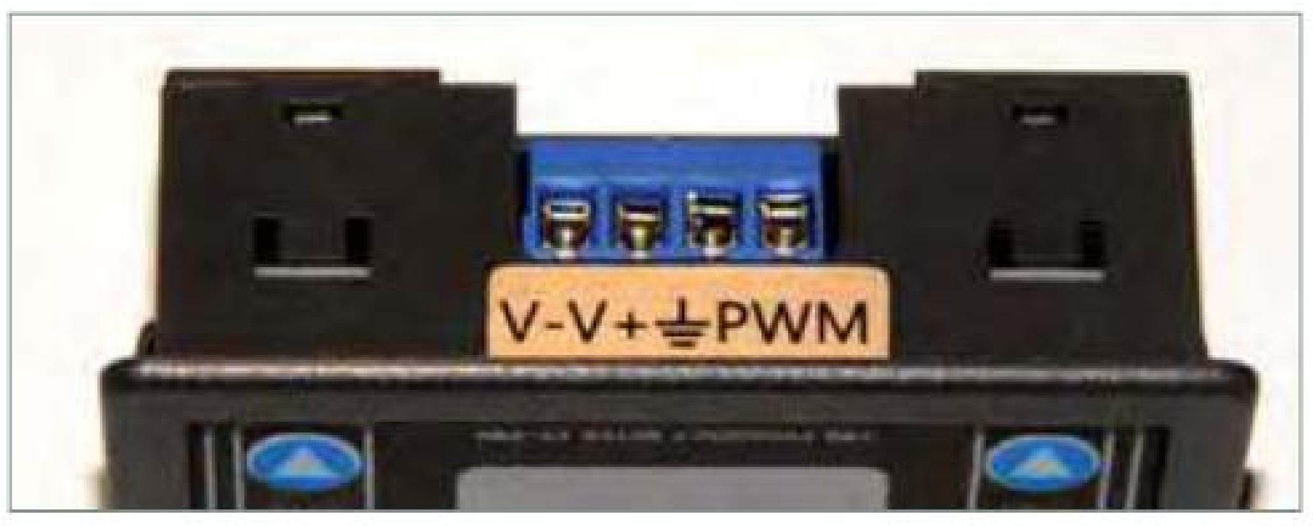
The frequency generator module has an in-built LCD that shows selected frequency and duty cycle. The up/down keys each side of the display allow you to select frequency (1Hz – 150kHz) and duty cycle (0-100%). Cost of the module is just over £4 delivered!

FREQ
OUT
350
040
%
DUTY
PWM FREQUENCY METER XY-PWM

The backlit LCD is clear and bright. Here the output has been set to a frequency of 350Hz and a duty cycle of 40%.

To find the module, search eBay under 'Signal Generator Test Equipment PWM Pulse Frequency Adjustable Module', at the time of writing item 303100197720. The cost is incredibly low – from around £4.15 delivered. Connections are straightforward – two for power and two for the output.

If you connect an LED to the output (remember to equip it with a dropping resistor to suit the supply voltage), you immediately have an excellent teaching tool for demonstrating what 'variable duty cycle' and 'variable frequency' mean. Students just watch the LED and alter the settings! Nothing brings these concepts more to life than holding a fixed frequency of (say) 1Hz and varying the duty cycle, and then holding a fixed duty cycle (eg, 20%) and varying the frequency. To make such a teaching tool, mount the display and a bright LED in a plastic box, powered by a 9V battery and equipped with an on/off switch.



The four connections to the frequency generator are well-labelled – power (–) and (+), and the PWM output – also polarised.

Note that when changing the frequency (but not the duty cycle), the output can momentarily go to 100% duty cycle, irrespective of the actual duty cycle output that has been set. This could be important in some applications.

MOSFET power module

So, how to run higher current loads than the 30mA the frequency generator module can deliver?

I initially connected the module to a standard DC:DC Solid State Relay (SSR) but found that the SSRs I tried didn't like being operated at high frequencies. (And plus, SSRs are still relatively expensive.) I then bought a small MOSFET-based PWM control module – dubbed on eBay '3-20V MOSFET MOS Transistor Trigger Switch Driver Board PWM Control Module BDAU', at the time of writing item 163330158628. This costs under £2 delivered.

Two inputs on this board are for the PWM control signal – provided in this case by the frequency generator module. There are also inputs for power and ground, and the output to the device being driven. The MOSFET is an FQD-60N03L, whose data sheet shows that it can handle quite high current (over 40A); however, that's obviously with plenty of heatsinking. In this application, the board's maker claims 10A, but I was happier with 5A continuous – this resulted in a 47°C running temperature in 20°C ambient conditions. (Note that the board can therefore drive

Using this MOSFET module allows the frequency generator to drive higher current loads – up to 5A continuous, and proportionally more at lower duty cycles.

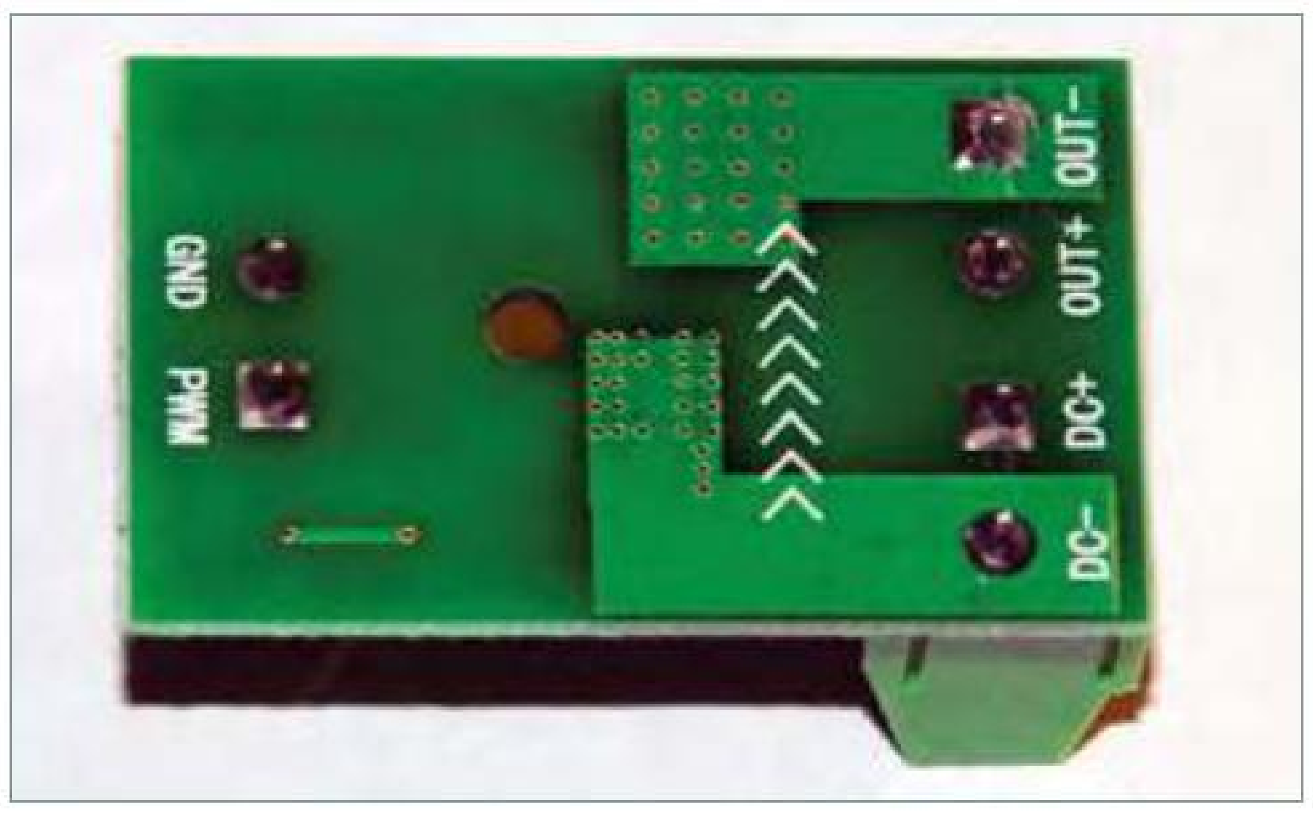
higher current loads at lower duty cycles – eg, 10A at 50% duty cycle.)

A maximum PWM frequency of 20kHz is specified. Any voltage from 3-20V can be used as the PWM signal, and the output will work over a 3.7-27V range. That makes this module a good match with the frequency generator – eg, run them off the same 4-20V supply.

Uses

The combination of the driver board and the variable frequency/duty cycle module allows you to do lots of things!

One use is testing 12V PWM solenoids, used as flow control valves. In addition to being switched on or off, solenoid valves can be triggered with a variable PWM signal, resulting in a variable flow capability. However, the duty cycle range over which a particular valve is effective, and the best frequency at which to pulse it, can often be found only through testing the valve. This set-up can easily achieve that.



As with the frequency generator, the connections are clearly labelled. The Gnd and PWM terminals connect to the respective terminals on the frequency generator. Power and ground connect to the (DC+) and (DC–) terminals, and the load is connected to the Out terminals.

Another application of a variable duty cycle / variable frequency driver is when testing car fuel injectors. When injectors are being flow tested, they need to be pulsed at frequencies that replicate how they are run in an engine (eg, 8-50Hz) and at different duty cycles, eg, from 3-80%. Again, these control modules can easily achieve that.

You can also use the system as a speed control for an electric motor – run from a regulated supply, you should be able to quite precisely and repeatedly dial-up specific speeds of devices like engravers, fans... whatever you need to drive. The system can also work as a low-voltage DC light dimmer – set a suitably high frequency and then control the brightness just by altering the duty cycle.

Finally, the system can be used to pulse a light or buzzer. The versatility of the settings means you can specify whatever frequency and duty cycle you want. For example, to save power, use a higher frequency (eg, 5Hz) and a lower duty cycle (eg, 10%). For maximum attention, use a lower frequency (eg, 1Hz) and a longer duty cycle (eg, 50%).

And I am sure that I've only just scratched the surface of possible uses!

Next month

In June, a slight change from my usual *Electronic Building Blocks* column. I'll be covering the use of thermal imaging cameras and asking are they worth it in hobby electronics? One thing I can vouch for – it took great images of Victor, our pet sheep, on a cold morning!

Teach-In 8 CD-ROM
Exploring the Arduino

This CD-ROM version of the exciting and popular *Teach-In 8* series has been designed for electronics enthusiasts who want to get to grips with the inexpensive, immensely popular Arduino microcontroller, as well as coding enthusiasts who want to explore hardware and interfacing. *Teach-In 8* provides a one-stop source of ideas and practical information.

The Arduino offers a remarkably effective platform for developing a huge variety of projects; from operating a set of Christmas tree lights to remotely controlling a robotic vehicle wirelessly or via the Internet. *Teach-In 8* is based around a series of practical projects with plenty of information for customisation. The projects can be combined together in many different ways in order to build more complex systems that can be used to solve a wide variety of home automation and environmental monitoring problems. The series includes topics such as RF technology, wireless networking and remote web access.

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PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, plated-through hole, with solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

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Back issues of articles are available – see Back Issues page for details.

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For editorial contact details see page 7.

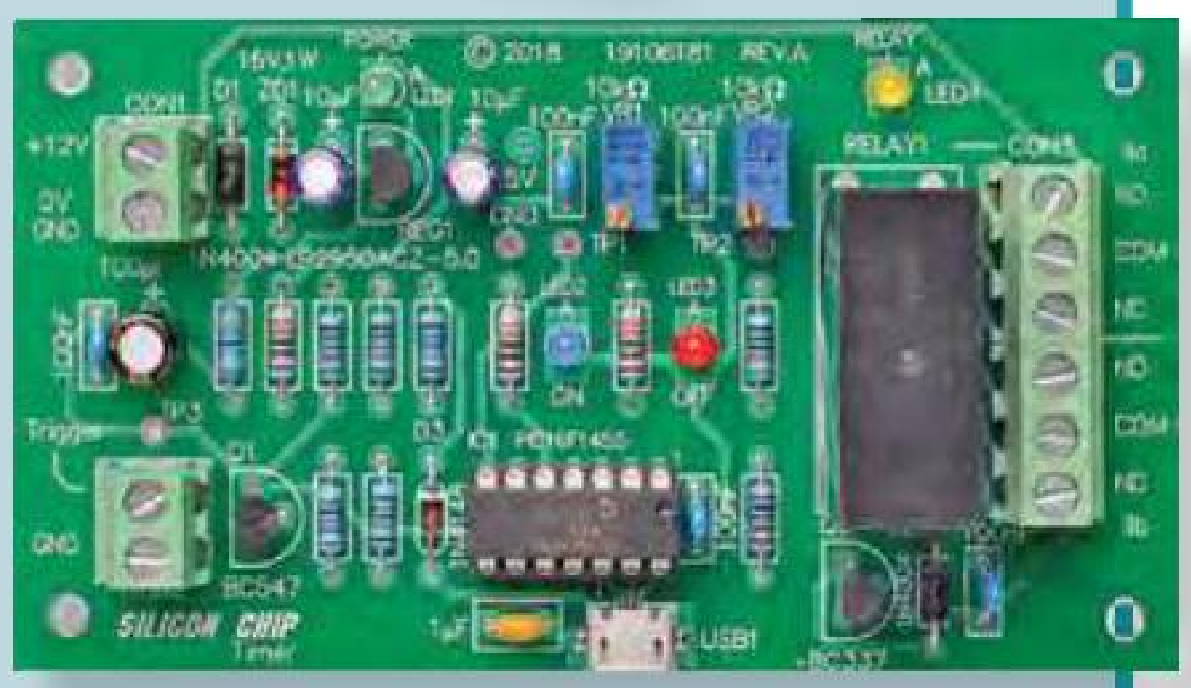
Next Month – in the June issue

LC Meter

This self-calibrating digital meter is based on a custom Arduino shield. It has an amazing measurement range from under 1pF to over 1F (yes, one farad!) and under 100nH to several henries. A must-build piece of equipment!

USB Flexitimer

Here's a very flexible timer that can switch its relay at intervals from milliseconds to days. It can be triggered by an external pulse or set to run continuously, switching its relay on and off at a particular interval.



MIDI Ultimate Synthesiser – Part 5

Next month is all about MIDI – both as a general topic and how it is implemented within the *MIDI Ultimate*.

Low-cost Electronic Modules – Part 17

Learn to use a highly linear 4GHz digital attenuator, which is programmable over 0-31.5dB in 0.5dB steps. Great for avoiding overload in a mixer or amplifier.

Teach-In 2019 – Part 7

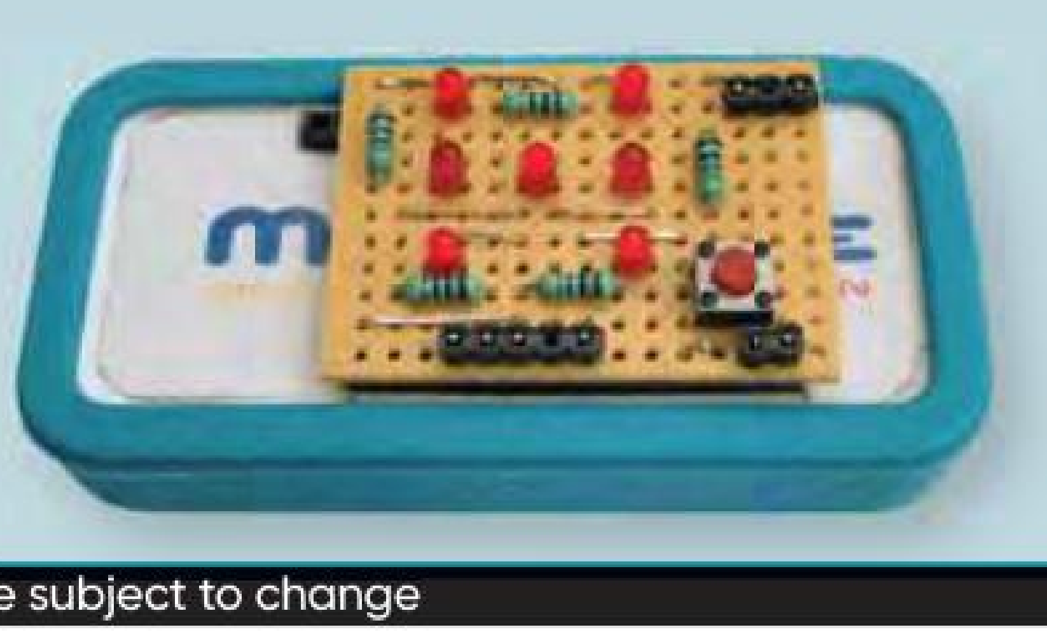
In *Part 7* of *Teach-In 2019* we will be introducing negative voltage converters and taking a detailed look at thermal design considerations for power supplies.

Make it with Micromite – Part 5

In June, more on digital inputs and outputs – we'll build the *Electronic Dice* mini-project.

PLUS!

All your favourite regular columns from *Cool Beans*, *Circuit Surgery* and *Electronic Building Blocks* to *PIC n' Mix*, *Audio Out*, *Techo Talk* and *Net Work*.



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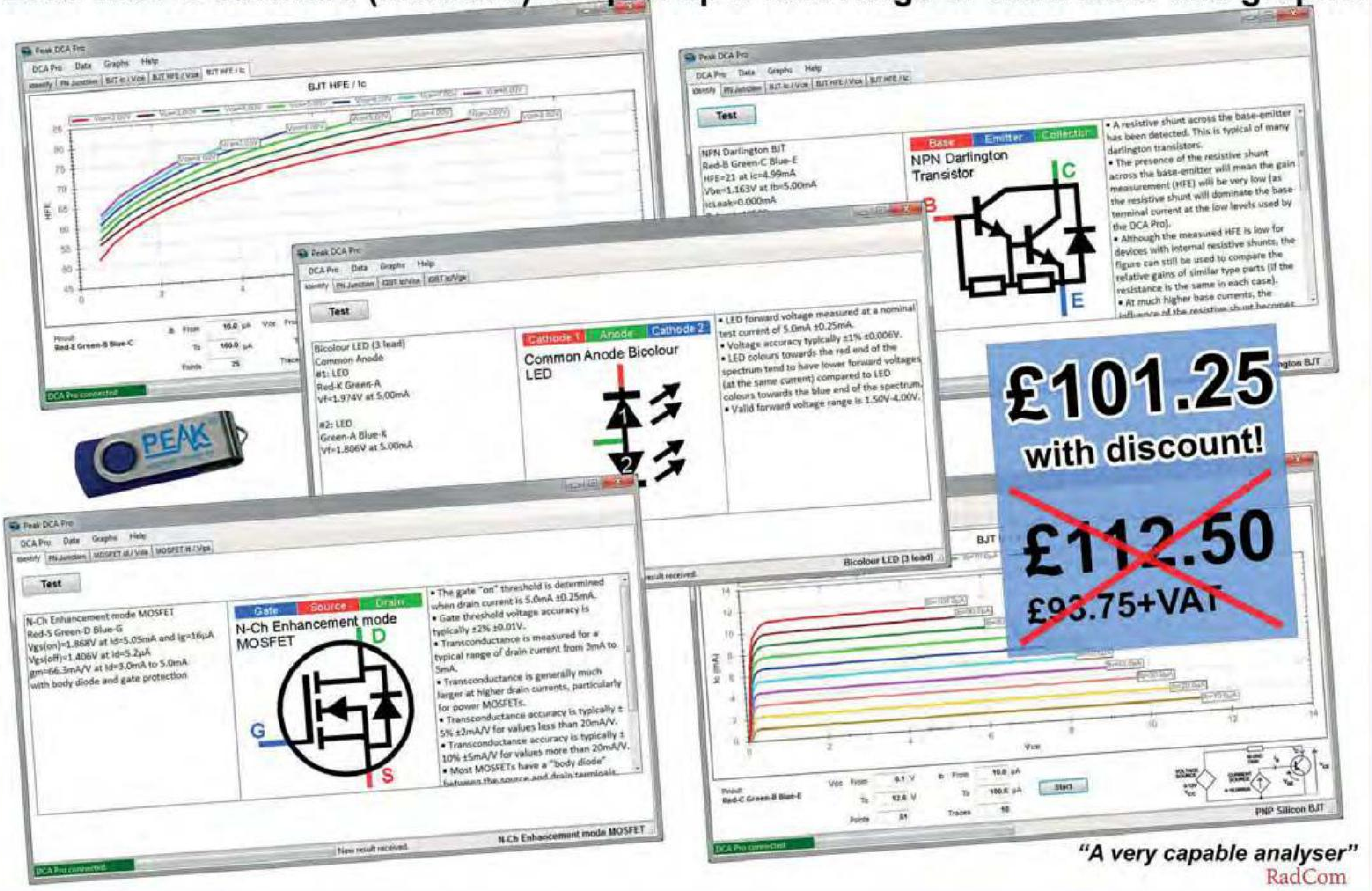
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Cathode 1
Anode
Cathode 2
Common Anode Bicolour LED
1
2
Gate
Source
Drain
N-Ch Enhancement mode MOSFET
G
D
S
PEAK
Vce
Ic (mA)
Start
PNP Silicon BJT
DCA Pro connected
New result received.
Bicolour LED (3 lead)
N-Ch Enhancement mode MOSFET

NPN Darlington BJT
Red-B Green-C Blue-E
HFE=21 at Ic=4.99mA
Vbe=1.163V at Ib=5.00mA
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- A resistive shunt across the base-emitter has been detected. This is typical of many darlington transistors.
- The presence of the resistive shunt across the base-emitter will mean the gain measurement (HFE) will be very low (as the resistive shunt will dominate the base terminal current at the low levels used by the DCA Pro).
- Although the measured HFE is low for devices with internal resistive shunts, the figure can still be used to compare the relative gains of similar type parts (if the resistance is the same in each case).
- At much higher base currents, the

Bicolour LED (3 lead)
Common Anode
#1: LED
Red-K Green-A
Vf=1.974V at 5.00mA

#2: LED
Green-A Blue-K
Vf=1.806V at 5.00mA

- LED forward voltage measured at a nominal test current of 5.0mA ±0.25mA.
- Voltage accuracy typically ±1% ±0.006V.
- LED colours towards the red end of the spectrum tend to have lower forward voltages (at the same current) compared to LED colours towards the blue end of the spectrum.
- Valid forward voltage range is 1.50V-4.00V.

N-Ch Enhancement mode MOSFET
Red-S Green-D Blue-G
Vgs(on)=1.868V at Id=5.05mA and Ig=16µA
Vgs(off)=1.406V at Id=5.2µA
gm=66.3mA/V at Id=3.0mA to 5.0mA
with body diode and gate protection

- The gate "on" threshold is determined when drain current is 5.0mA ±0.25mA.
- Gate threshold voltage accuracy is typically ±2% ±0.01V.
- Transconductance is measured for a typical range of drain current from 3mA to 5mA.
- Transconductance is generally much larger at higher drain currents, particularly for power MOSFETs.
- Transconductance accuracy is typically ± 5% ±2mA/V for values less than 20mA/V.
- Transconductance accuracy is typically ± 10% ±5mA/V for values more than 20mA/V.
- Most MOSFETs have a "body diode"

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